# COMMUNICATIONS INTERFERENCE

# REDUCTION STUDY

by

W. B. Warren, Jr.

Project A-525

Contract No. AF 30(602)-2399

Prepared for
Rome Air Development Center
Air Force Systems Command
United States Air Force
Griffiss Air Force Base
New York

Engineering Experiment Station
Georgia Institute of Technology
Atlanta
1960-62

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**GEORGIA INSTITUTE OF TECHNOLOGY**
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

5 December 1960

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

ATTENTION: RCUAC

SUBJECT: Monthly Progress Letter No. 1, Contract No. AF 30(602)-2366

Gentlemen:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

Preliminary work has been done on the design of a limiter using a controlled threshold which can be set low enough to give effective limiting action at the very low signal level encountered at the input to a communications receiver. Several possibilities have been considered. They are:

1. Non-linear Capacitor in Parallel with a Resonant Circuit

The sketch of Figure 1 indicates a method in which a voltage sensitive capacitor was connected in parallel with a resonant circuit. In this arrangement, the positive peaks of the signal appearing across the tuned circuit caused the diode to conduct, charging the fixed capacitor to a voltage equal to the peak of the applied signal. When the signal swings in the negative direction the diode is back-biased and exhibits capacitance which is a function of the dc voltage across the fixed capacitance. Since this dc voltage is equal to the peak of the input signal the diode capacitance will vary with the amplitude of the applied signal. If the resonant circuit is initially tuned to resonance for small signals, then at large signal levels the bias voltage developed across the fixed capacitance will cause the resonant circuit to be de-tuned, and the gain to the applied signal will be reduced.

When a large amplitude signal was applied to the circuit of Figure 1, the desired limiting action was present when the frequency of the applied signal was below the resonant frequency of the circuit. However, when the frequency of the applied signal was on the opposite side of resonance, an unstable envelope in the output was obtained. This was due to the fact that as the amplitude of the applied signal was increased, the bias voltage increased, tuning the circuit closer to the frequency of the applied signal. This resulted in an even larger bias voltage and subsequent tuning of the circuit even closer to resonance. This action is a cumulative one, so that the envelope becomes unstable as shown in the sketch of Figure 2. This effect is undesirable in obtaining limiting action since whether or not limiting is obtained depends upon the frequency of the applied signal.

2. Negative and Positive Resistors in Parallel

If a positive resistor is connected in parallel with a tunnel diode possessing a volt-ampere characteristic such as that shown in Figure 3, the overall volt-ampere characteristic of the combination will appear as shown in the composite curve of Figure 3, provided the value of the positive resistor is properly selected. Inspection of the voltage scale shows that the width of the flat portion of the composite curve, in volts, is quite small. Since the resistance represented by the curve at any point is the reciprocal of the slope of the curve, this flat portion represents a very high resistance, while those portions of the curve on either side of this flat portion represent a relatively low resistance. Thus, by biasing the parallel combination of the positive resistor and the tunnel diode in the center of the flat region, it should be possible to obtain excellent limiting action on those signals whose amplitudes exceed one-half the width of the flat portion of the curve.

3. Oscillating Parametric Amplifier

It is known that a parametric lower sideband up converter is potentially unstable, while an upper sideband up converter is absolutely stable. It is possible to make use of this characteristic of parametric converters to obtain limiting action. If the pump frequency is considerably higher than the signal frequency and if the desired signal output is taken at the upper sideband frequency while the output at the lower sideband frequency is terminated in a resistor, then the input-output characteristic of a desired signal is linear, provided the pump and signal levels are sufficiently low. However, if sufficient signal or pump power is supplied to cause oscillation at the lower sideband frequency to take place, then the signal and pump energy will be used to maintain the oscillation, and the gain for the desired signal will be reduced. Thus, if the pump level is set just below the value required to cause oscillation, a very small input signal is sufficient to cause the start of oscillation and a limiting characteristic for the input signal is obtained.

During the next month, most of the effort will be expended on investigation of the limiting parametric amplifier.

Finances:

The amount of funds remaining in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

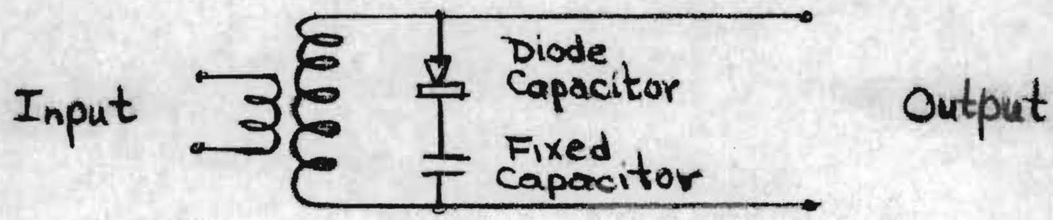



Figure 1-Resonant Circuit With Non-Linear Capacitance

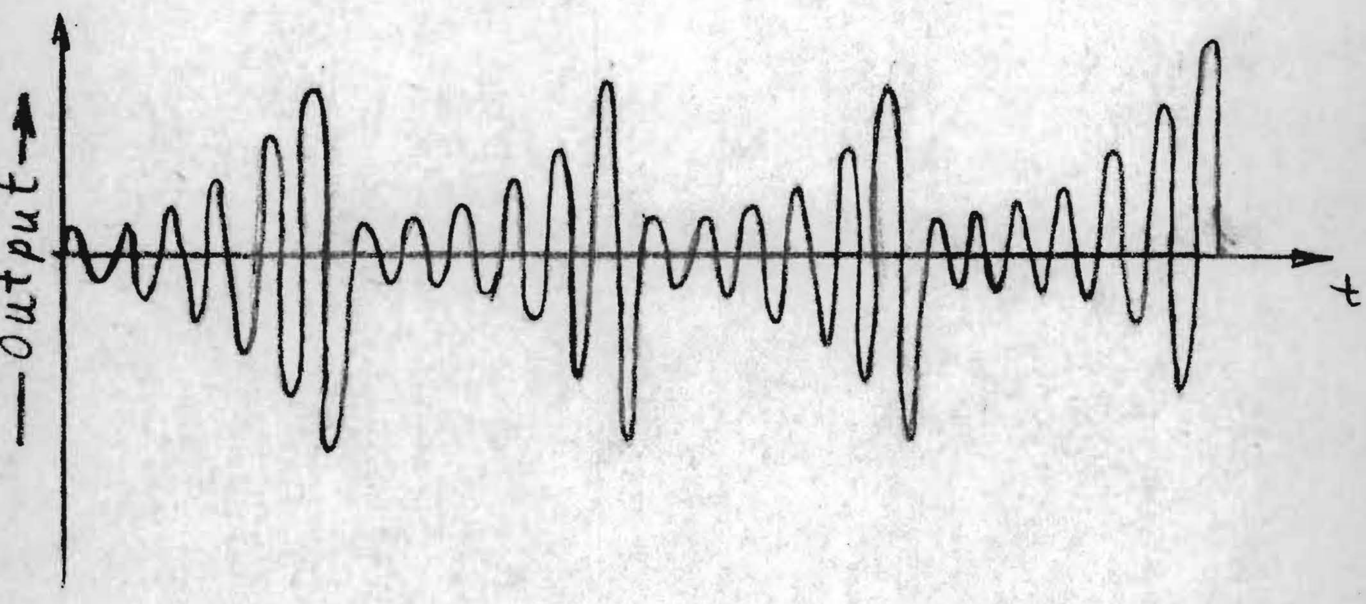



Figure 2- Envelope Instability

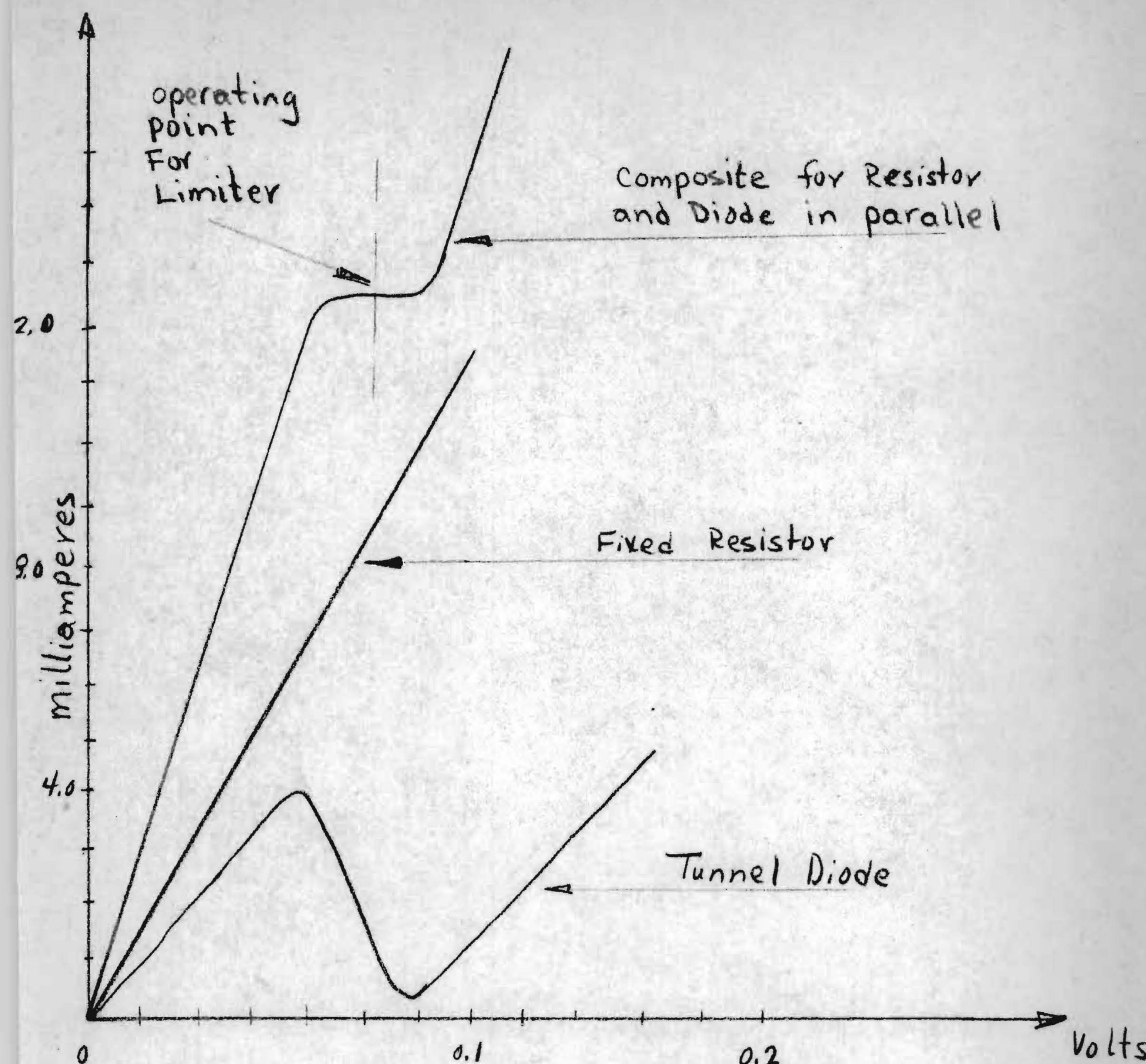



Figure 3- Resistor and Tunnel Diode in Parallel

**GEORGIA INSTITUTE OF TECHNOLOGY**
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

13 January 1961

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

ATTENTION: RCUAC

SUBJECT: Monthly Progress Letter No. 2, Contract No. AF 30(602)-2366

Gentlemen:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

During December, the work on construction of a limiting parametric amplifier was curtailed in order that a more thorough literature search might be made before undertaking the construction of this device. Several papers dealing with the mechanism of limiting in parametric systems were examined, but most of these dealt with limiting at high power levels. However, some papers did report successful operation at low signal levels with one report of successful limiting at a signal level of -40 dbm. As a result, construction efforts were directed toward the packaging of an auxiliary unit which would perform the function of interference suppression by either the sampling or blanking technique. This unit includes provision for synchronizing the internal sampling or blanking pulse generator with an incoming interfering pulse signal. This sampling or blanking pulse generator is used to switch a blanker connected in the antenna lead of the receiver in the proper fashion to exclude the interference from the receiver input. This unit requires no internal connection or modification of the receiver with which it is being used. Construction of this device is approximately 50% complete.

In the next month, construction of this unit will be continued. In addition, construction of the limiting parametric amplifier will be undertaken.

Finance:

The amount of funds remaining in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

**GEORGIA INSTITUTE OF TECHNOLOGY**
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

7 February 1961

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

ATTENTION: RCUAC

SUBJECT: Monthly Progress Letter No. 3, Contract No. AF 30(602)-2366

Gentlemen:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

During the previous month, the major effort was placed on obtaining a breadboard model of a limiting parametric amplifier. Such an amplifier has been constructed, and successful limiting at a signal level of approximately 0 dbm has been obtained. In this device the limiting action is obtained by forcing the parametric device into oscillation at a sub-multiple of the input frequency. This causes the input signal energy to be used up in the maintenance of the oscillation at the sub-harmonic frequency. At low signal levels the oscillation takes place at half frequency but, as the input signal is increased in amplitude, the order of the sub-harmonic increases. However, the output of the device is taken at the same frequency as the input signal, and it is this transfer which exhibits the limiting action.

The use of a sub-harmonic mode rather than the more conventional parametric amplifier which uses an idler frequency has the advantage that fewer components are required. This is due to the fact that the tuned circuit at one-half the input frequency acts as its own idler. The construction of a low frequency device is illustrated in the sketch of Figure 1. The frequencies involved were chosen so that conventional test equipment might be used in evaluating the device.

Continuing improvement of this limiting technique will be carried out during February. In addition, work will be started on the design of an audio filter to reject periodic interference.

Finance:

The amount of funds remaining in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

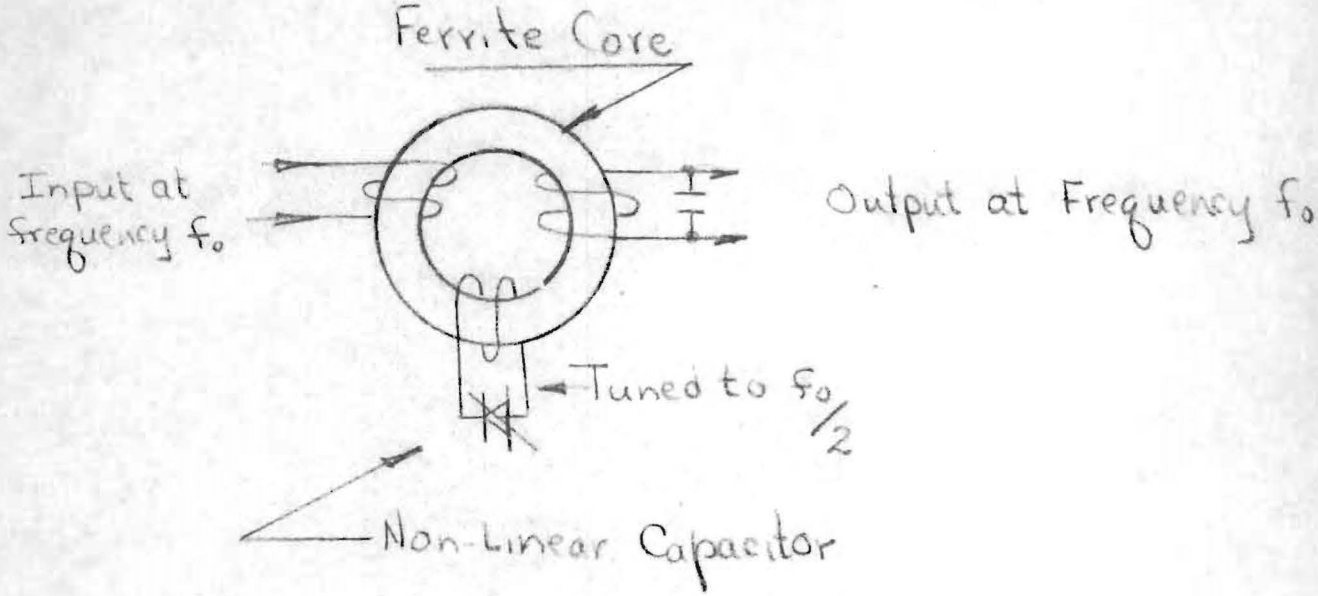



Figure 1 - Parametric Limiter

# GEORGIA INSTITUTE OF TECHNOLOGY

**ENGINEERING EXPERIMENT STATION**

**ATLANTA 13, GEORGIA**

16 March 1961

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

ATTENTION: RCUAC

SUBJECT: Monthly Progress Letter No. 4, Contract No. AF 30(602)-2366

A-525

Gentlemen:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

Work on the auxiliary unit to provide pulse suppression by blanking has continued during the month of February. Some difficulty has been experienced in obtaining proper synchronization of the blanking rate at the proper multiple of the interferring pulse rate. During March efforts will be made to overcome this difficulty.

Preliminary design work has been completed on the audio filter to reject periodic interference. The variable delay line has been ordered and further work on the audio filter has been curtailed until the delay line is received.

A parametric limiter has been constructed in breadboard fashion to operate at 250 mc. Limiting action similar to that obtained at lower frequencies has been observed. This action takes place due to the input energy being dissipated at one half the input frequency. Hence the limiting threshold occurs at the point where the oscillation at the half frequency commences. For the breadboard model constructed, this threshold is about 0 dbm. The circuit configuration of this limiter is shown in the attached figure.

Finance:

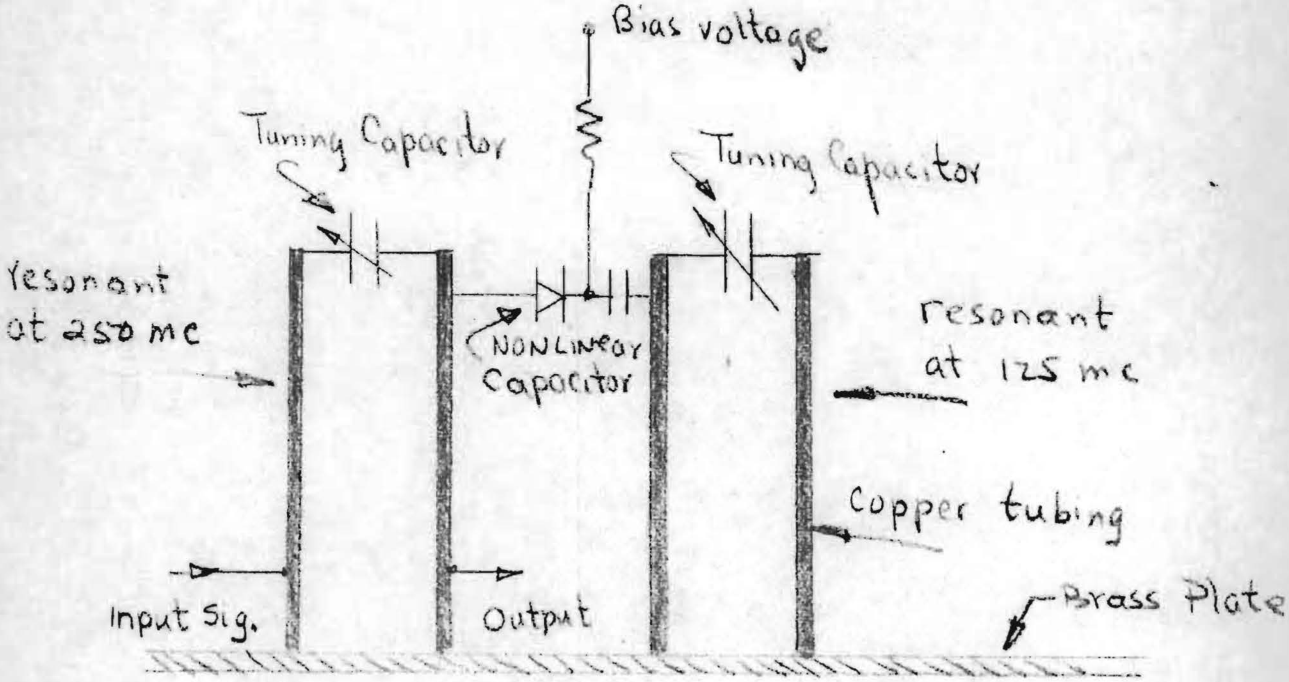
The amount of funds remaining in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch




Parametric Limiter

# GEORGIA INSTITUTE OF TECHNOLOGY

ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

14 April 1961

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

ATTENTION: RCUMA

SUBJECT: Monthly Progress Letter No. 5, Contract No. AF 30(602)-2366

Gentlemen:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

During March, progress was made in the analysis of the response of resonant circuits containing nonlinear capacitance. In particular, the conditions for the start of oscillation in the circuit of Figure 1 were obtained. This is the circuit configuration being used to obtain limiting action at high signal levels. The signal level at the start of the half-frequency oscillation is small, the linear term in the variation of capacitance with voltage is used, since the effects of higher order terms become smaller and smaller as the signal level tends toward zero. Therefore the variation of capacitance with voltage is given by:

$$C(V_c) = C_o + C_1 V_c . \tag{1}$$

From the definition of capacitance,

$$C(V_c) = \frac{Q_c}{V_c} , \tag{2}$$

or

$$Q_c = V_c \, C(V_c).$$

But

$$i_c = \frac{dQ_c}{dt} = \frac{d}{dt}\left( C_oV_c + C_1V_c^{\,2}\right) = \left( C_o + 2C_1V_c\right)\frac{dV_c}{dt} . \qquad (3)$$

From Figure 1, the voltage across the nonlinear capacitor is

$$V_c = (V_i - V_o). \qquad (4)$$

If $V_i = A \cos \omega_1 t$,

and $V_o$ is assumed to be of the form:

$$V_o = P \cos \frac{\omega_1 t}{2} + Q \sin \frac{\omega_1 t}{2} \qquad (5)$$

(since the impedance of the tuned circuit is small at all frequencies except $\omega = \frac{\omega_1}{2}$ , components of $V_o$ at other frequencies will be small), then, from (4),

$$V_c = \left( A \cos \omega_1 t - P \cos \frac{\omega t}{2} - Q \sin \frac{\omega t}{2} \right) . \qquad (6)$$

If (6) is substituted in (3), the indicated operations performed, and components at frequency $\frac{\omega_1}{2}$ are collected, there results:

$$i\Big|_{\frac{\omega_1}{2}} = \left(\frac{C_1 \; P \; A\omega_1}{2}\right) \sin \frac{\omega_1 t}{2} + \left(\frac{C_1 \; Q \; A\omega_1}{2}\right) \cos \frac{\omega_1 t}{2} . \qquad (7)$$

But $$V_o\Big|_{\omega=\frac{\omega_1}{2}} = \left( i\Big|_{\frac{\omega_1}{2}} \right) \left( Z_t\Big|_{\frac{\omega_1}{2}} \right), \qquad (8)$$

where $Z_t\Big|_{\frac{\omega_1}{2}}$ is the impedance of the resonant output circuit evaluated at $\omega = \frac{\omega_1}{2}$. This is equal to R for the circuit of Figure 1.

Therefore

$$V_o = \left(\frac{C_1 \; P \; A\omega_1 \; R}{2}\right) \sin \frac{\omega_1 t}{2} + \left(\frac{C_1 \; Q \; A\omega_1 R}{2}\right) \cos \frac{\omega_1 t}{2} . \qquad (9)$$

But $V_o$ was assumed to be

$$V_o = P \cos \frac{\omega_1 t}{2} + Q \sin \frac{\omega_1 t}{2} \qquad (5)$$

Therefore:

$$\left.\begin{aligned} &\frac{R \; C_1 \; P \; A\omega_1}{2} = Q, \text{ and} \\ &\frac{R \; C_1 \; A \; Q\omega_1}{2} = P \text{ simultaneously.} \end{aligned}\right\} \qquad (10)$$

For P and $Q \neq 0$, solutions are

$$\left.\begin{aligned} &Q = \pm P, \text{ and} \\ &A = \frac{2}{\omega_1 C_1 R} . \end{aligned}\right\} \qquad (11)$$

This says that the amplitude of the input signal must exceed the value $\frac{2}{\omega_1 C_1 R}$ for oscillation, and hence limiting, to begin; and furthermore the phase angle of this oscillation will be $\pm 45^{\circ}$ if the output circuit is tuned to $\omega = \frac{\omega_1}{2}$.

These conditions are independent of both P and Q and hence give no information concerning the amplitude of the half frequency oscillation. This is the result of assuming only linear variation in the capacitance with voltage. Further analysis will be pointed toward the inclusion of higher order terms in order to obtain information concerning the exact shape of the limiting characteristic.

The current state of the blanking and sampling adaptor is shown in the block diagram of Figure 2. Two more sections have been added to the phase shifter to increase the range of phase adjustment to greater than $360^{\circ}$, thereby providing any phase shift necessary for either blanking or sampling. The phase shift circuitry being used is shown in Figure 3.

Preliminary work is being done on the receiver and toward packaging of the entire adaptor in a form suitable for use with a conventional receiver.

Finances:

The amount of funds remaining in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

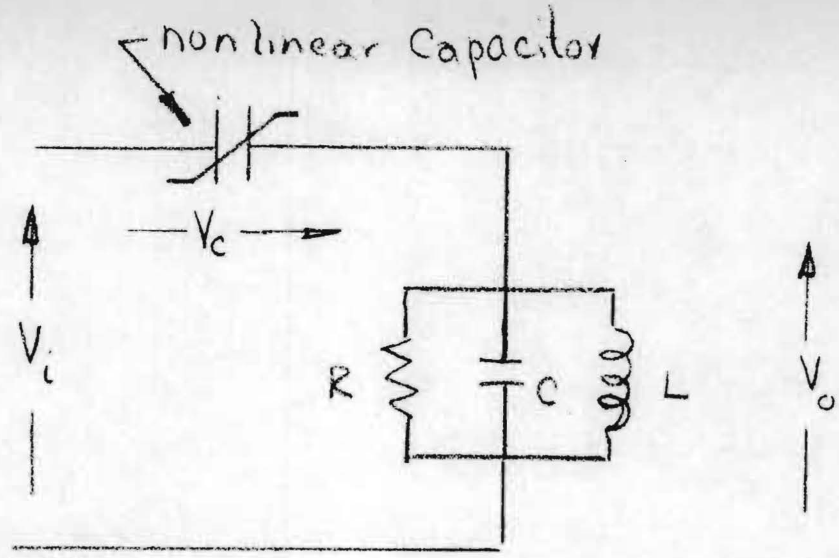



Figure 1 - Nonlinear Circuit

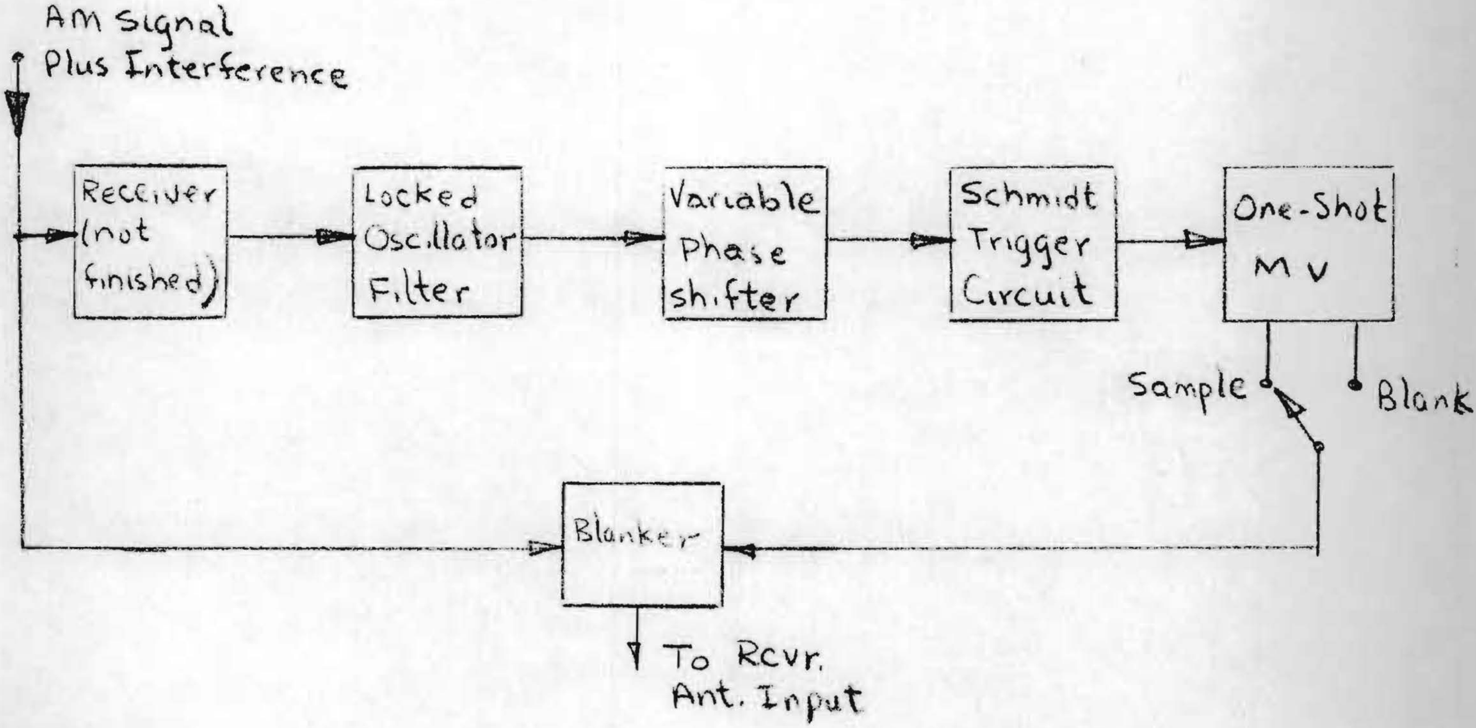



Figure 2 - Receiver Adaptor

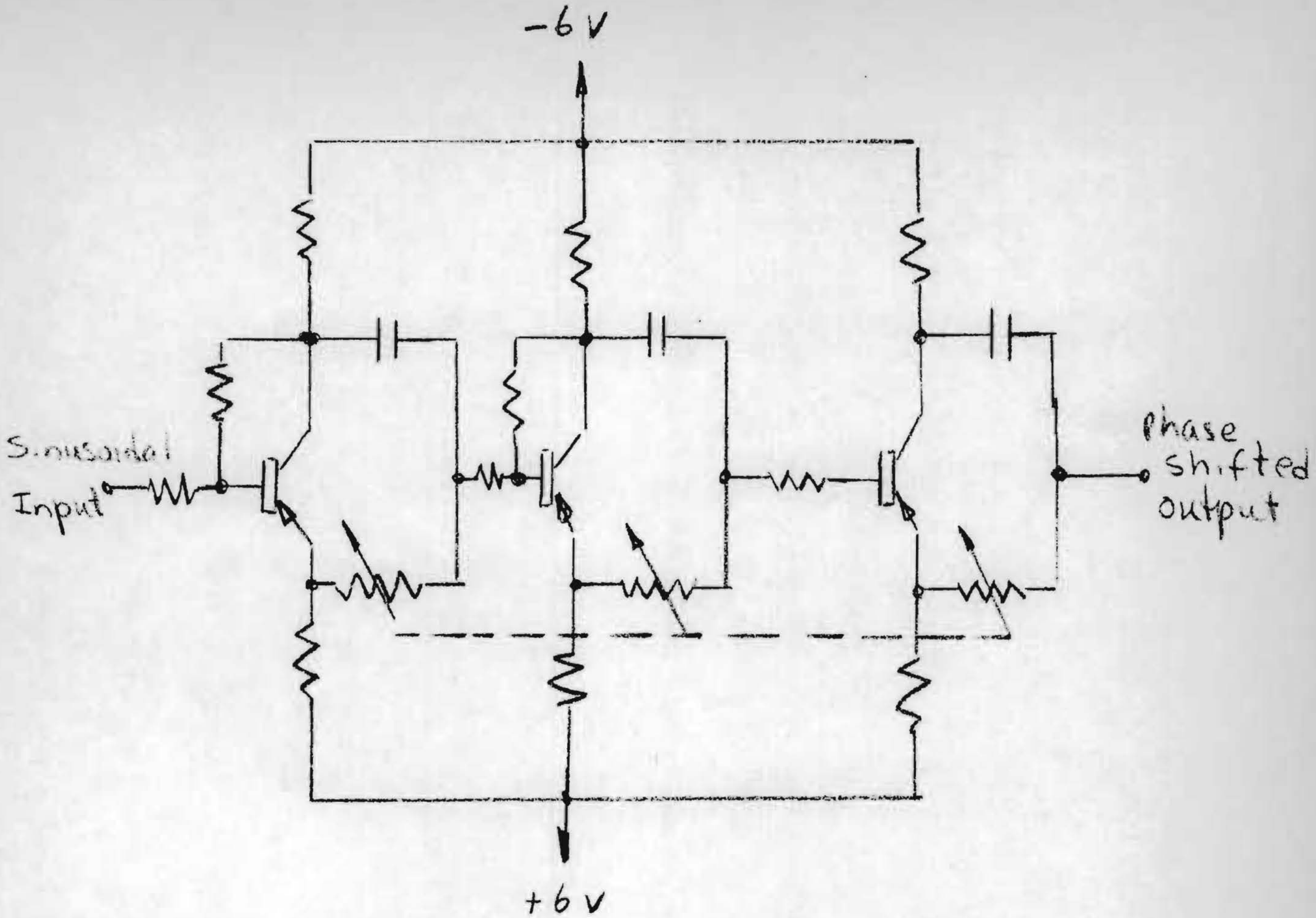



Figure 3 - Variable Phase Shifter

GEORGIA INSTITUTE OF TECHNOLOGY
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

17 May 1961

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

ATTENTION: RCUMA

SUBJECT: Monthly Progress Letter No. 6, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

Project efforts during April were concentrated mainly upon improvement of the blanking adaptor equipment. With the exception of the auxiliary receiver, the design work is now complete and construction of the adaptor in a form suitable for demonstration has been started. Final design improvements made during April consisted primarily of improvements in the locked oscillator filter which extracts the proper harmonic of the interference pulse rate so as to provide a blanking rate which is harmonically related to the interference rate. The schematic of Figure 1 shows the method by which a one-shot multivibrator has been added in front of the locked oscillator to standardize the synchronizing signal. This addition makes the synchronization of the oscillator insensitive to the shape of the synchronizing pulses out of the auxiliary receiver. The injection of the synchronizing signal into the oscillator is by means of a two turn link on the oscillator coil so that the injected voltage in the oscillator tank circuit is proportional to the current in the left-hand transistor of the multivibrator. The start of this current is initiated by the input signal from the receiver, but the duration and amplitude are controlled by the constants of the multivibrator. Hence, the magnitude of the injection voltage of the oscillator is independent of the shape of the synchronizing signal from the receiver.

Efforts were made during April to extend the analysis of the parametric limiter in the manner described in the progress letter for March, but no satisfactory technique has been devised to assess the effects of higher order nonlinearities on the limiting action. Further efforts on the analysis of this problem will be directed toward the adaptation of a method used by Leeson[1] in the analysis of nonlinear capacitors as harmonic generators.

The delay line for use in the audio filter has been received and construction of this device will be started in May.

Finances:

The amount of funds remaining in the contract is sufficient to cover the present and anticipated rate of expenditure.

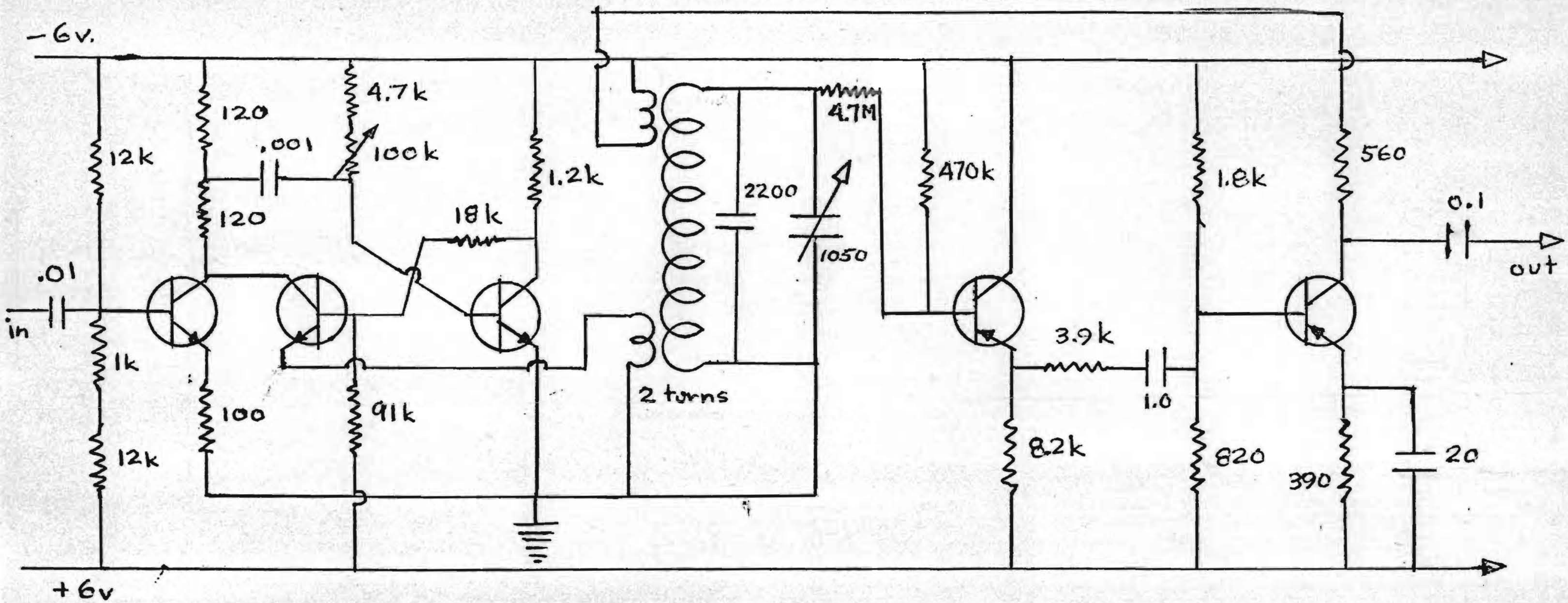
Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

for W. B. Wrigley, Head
Communications Branch

[1] Leeson, D.B. and Weinreb, S., "Frequency Multiplication with Nonlinear Capacitors," Proc IRE 47, pp 2076-2084, December 1959



-6v.
+6v
in
out
.01
12k
1k
12k
120
120
100
.001
4.7k
100k
91k
18k
1.2k
2 turns
2200
1050
4.7M
470k
8.2k
3.9k
1.0
1.8k
820
560
390
0.1
20


Fig 1

GEORGIA INSTITUTE OF TECHNOLOGY
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

14 June 1961

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

ATTENTION: RCUMA

SUBJECT: Monthly Progress Letter No. 7, Contract No. Af 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

During May, work was continued on the construction of the blanking adaptor. In particular, design has been completed for the auxilliary receiver. After several attempts to obtain sufficient sensitivity in this receiver, using a simple diode detector, it was concluded that a superheterodyne type of receiver will be necessary to obtain adequate sensitivity. This is necessary since at very low signal levels the nonlinearity of the simple diode detector is insufficient to give adequate rectification efficiency. The receiver now being constructed is outlined in the block diagram of Figure 1. The 40 Mc IF frequency was chosen to permit adequate band width to be obtained with a simple synchronously tuned IF strip. In addition, a power supply is being constructed to permit ac operation of the entire equipment.

A delay line filter for the rejection of periodic signals at low frequencies has been constructed in elementary form and its ability to reject these periodic signals has been determined. The method by which this filter was constructed is shown in the block diagram of Figure 2. The configuration is a simple cancellation scheme by which the phase shift through the delay line is some even multiple of 180° at the fundamental frequency and at each of the harmonics of the periodic signal.

Referring to Figure 2, the input signal is:

$$e_{in} = e_1 (t) \qquad (1)$$

Then the delayed signal will be given by:

$$e_{delayed} = e_1(t - \tau) \tag{2}$$

and the output signal is:

$$e_o(t) = e_{delayed} = e_1(t) - e_1(t - \tau) \tag{3}$$

Taking the Laplace transform of both sides gives:

$$E_o(s) = E_1(s)e^{-s\tau} \tag{4}$$

or:

$$\frac{E_o(s)}{E_1(s)} = \left(1 - e^{-st}\right) \tag{5}$$

The steady state response is obtained by letting:

$$s = j\omega \tag{6}$$

Then:

$$\text{Gain } (j\omega) = \frac{E_o(j\omega)}{E_1(j\omega)} = \left(1 - e^{-j\omega\tau}\right) \tag{7}$$

The zeros of this gain function occur for:

$$e^{j\omega\tau} = \cos \omega\tau + j \sin\omega\tau = 1 + jo \tag{8}$$

or:

$$\omega\tau = 2N\pi \tag{9}$$

or:

$$\omega = \frac{2N\pi}{\tau} \tag{10}$$

These are the values of $\omega$ which the fundamental and harmonic frequencies of the periodic interference must have if it is to be rejected. A sketch of the gain function of equation (7) is shown in Figure 3.

Because the delay obtainable with the available delay line is only 20 microseconds, the fundamental frequency of the periodic signal to be rejected must be 50 kc rather than an audio frequency as desired. However, the operation at an audio frequency only requires an appropriately longer delay line, and the performance exhibited by the 20 microsecond delay line filter illustrates the performance to be expected from a filter designed for audio frequency. In the test performed, a 50 kc square wave test signal was used. When the test signal frequency was set to 45 kc, considerable output from the filter was obtained. When the test signal frequency was set to 50 kc, the level of the filter output was approximately 40 db below that obtained at 45 kc.

No further progress has been made during May concerning the inclusion of higher order nonlinearity in the solution of the parametric limiter problem. Efforts to obtain the solution will be continued during June.

Finances:

The amount of funds remaining in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.

Approved:

W. B. Wrigley, Head
Communications Branch

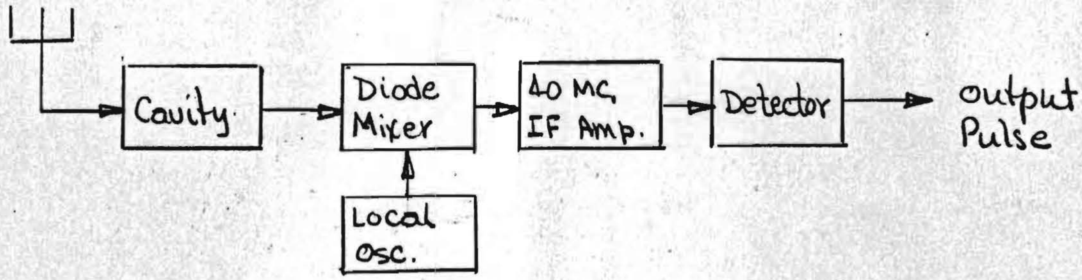



Figure 1. Auxilliary Receiver

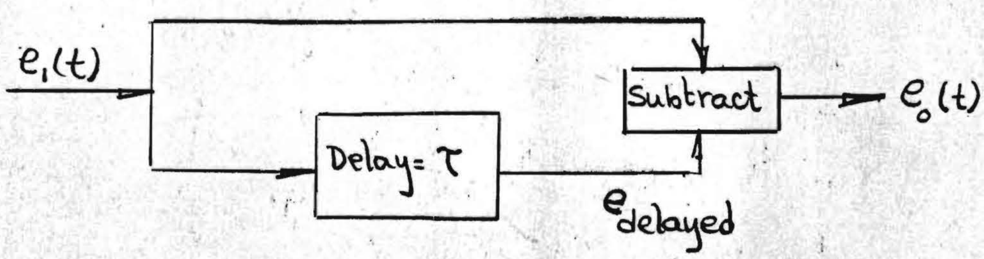



Figure 2. Delay Line Filter

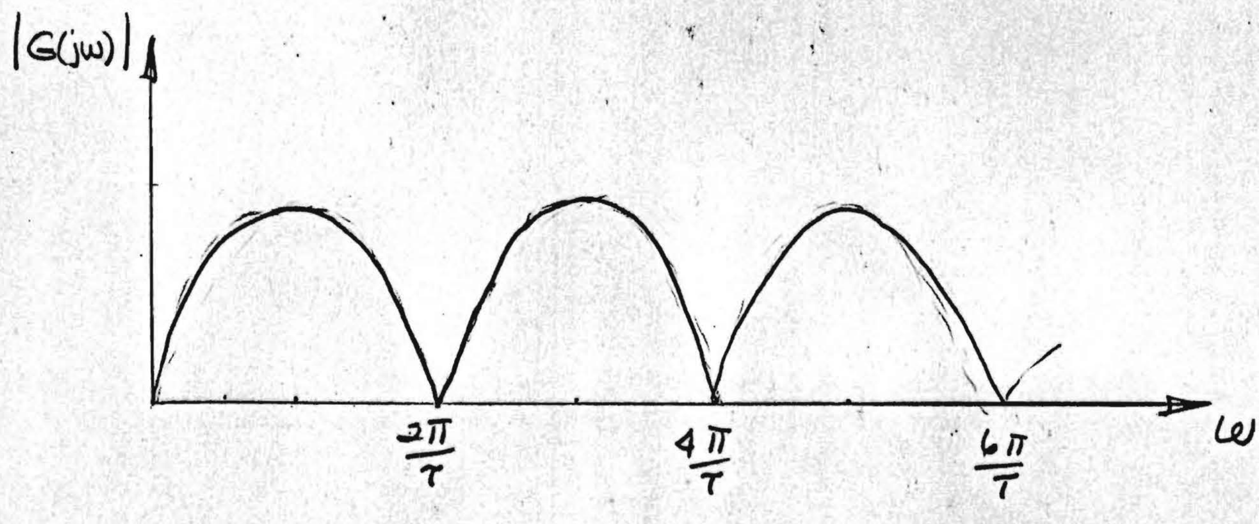



Figure 3. Response of Delay Line Filter

GEORGIA INSTITUTE OF TECHNOLOGY
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

18 July 1961

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

ATTENTION: RCUMA

SUBJECT: Monthly Progress Letter No. 8, Contract No. AF 30(602)-2366

A-525

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

Construction of the auxiliary receiver was completed during June, and a sensitivity of approximately 50 microvolts was measured, which considerably exceeds the required sensitivity of one millivolt.

Several networks have been investigated to determine a configuration suitable for use in the audio filter for periodic signals. A tunable null network is desired so that several of these networks can be connected in cascade to provide a transmission function with periodically spaced zeros. Each zero would be provided by one of the null networks, and each of the networks would be mechanically tracked with the others to provide a single tuning control. The arrangement is illustrated in the block diagram of Figure 1.

One null network which has been studied is shown in Figure 2. In this circuit, the transmission gain is given by:

$$\text{Gain} = \frac{E_o}{E_i} = \left( \frac{(1+j\omega CR_1)^2}{1-\omega^2C^2R_1{}^2+3j\omega CR_1} - \frac{R_3}{(R_2+R_3)} \right) \qquad (1)$$

This function has a zero at:

$$\begin{cases} \omega = \dfrac{1}{RC} \\ \dfrac{R_3}{R_2+R_3} = \dfrac{2}{3} \end{cases} \qquad (2)$$

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However, the tuning of this zero requires the use of a dual potentiometer at both the positions where $R_1$ appears. To overcome this drawback, another network was examined, having a null which may be tuned by a single potentiometer. The network is shown in Figure 3.

The gain function for this network is given by:

$$\text{Gain} = \frac{E_o}{E_i} = \frac{1 - K^2\omega^2R^2C^2}{(1 - \omega^2C^2R^2) + 3j\omega CR} \qquad (3)$$

This function has a zero at:

$$\omega = \frac{1}{KRC} \qquad (4)$$

The null frequency is seen to be inversely proportional to the attenuation, K. One practical realization of this network is shown in Figure 4.

During July, several of these networks will be constructed to determine the feasibility of using them to reject periodic interference.

Finances:

The amount of funds remaining in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

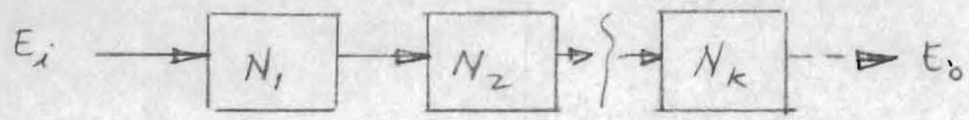


Figure 1- Cascade of Null Networks to Obtain Periodic Zeroes

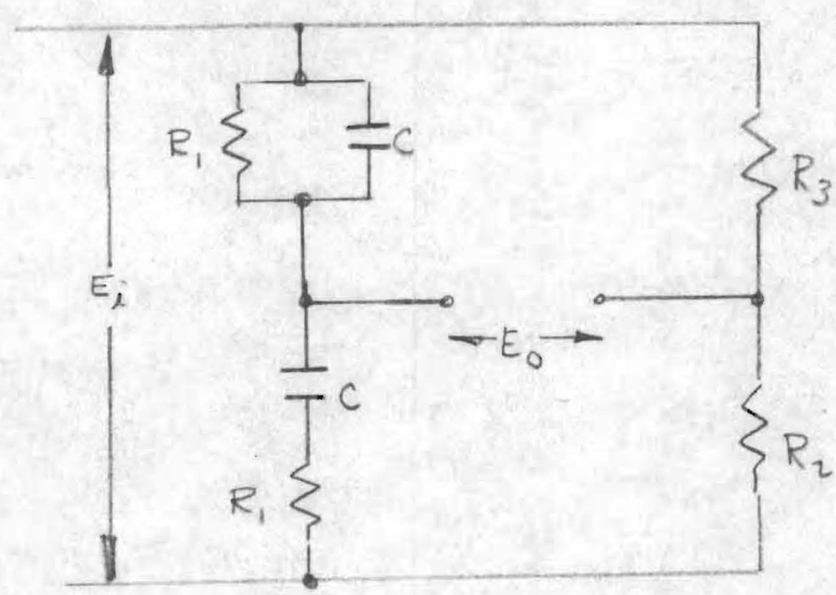


Figure 2 - Null Network

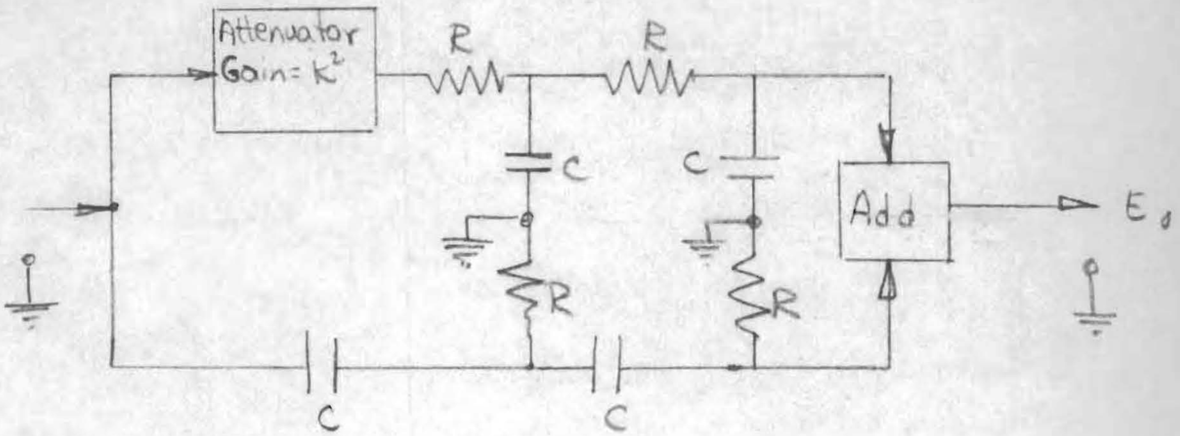


Figure 3 - Null Network

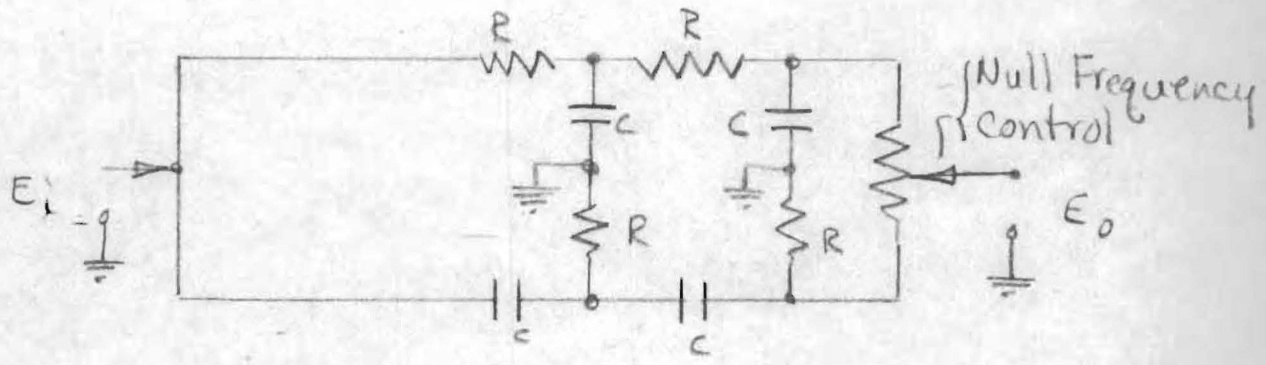


Figure 4 - Practical Null Network

GEORGIA INSTITUTE OF TECHNOLOGY
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

18 August 1961

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

ATTENTION: RCUMA

SUBJECT: Monthly Progress Letter No. 9, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

Electrical design and preliminary testing of the blanking adaptor is now complete. Efforts in July were directed toward a clean-up of details of mechanical construction and fabrication of a power supply. Tests of the completed unit will be performed in August.

Investigation of audio null networks to reject periodic interference was continued, with several new configurations being discovered, both in the literature and by project personnel. Of particular interest is the network shown in Figure 1. This network is the topological dual of the single control null network reported in the progress letter for June, and its current transfer function is given by:

$$\frac{I_{in}}{I_{out}} = \frac{1 - K^2\omega^2C^2R^2}{1 - \omega^2C^2R^2 + 3j\omega CR}$$

The transmission function of interest for this network is the short circuit current ratio rather than the open circuit voltage ratio. This form of the network is more useful for transistor circuitry because of the inherently low input impedance of transistors. A typical circuit using this null network is shown in Figure 2.

An extensive list of null networks has recently appeared in the literature[1], of which several appear to have some applicability to the work of this project. During August, several of these networks will be constructed to determine their effectiveness in the construction of audio filters with a periodic null response.

Finances:

The amount of funds remaining in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

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[1] "Single Component-Controlled RC Null Networks", General Radio Experimentor, Vol. 35, No. 7, July 1961.

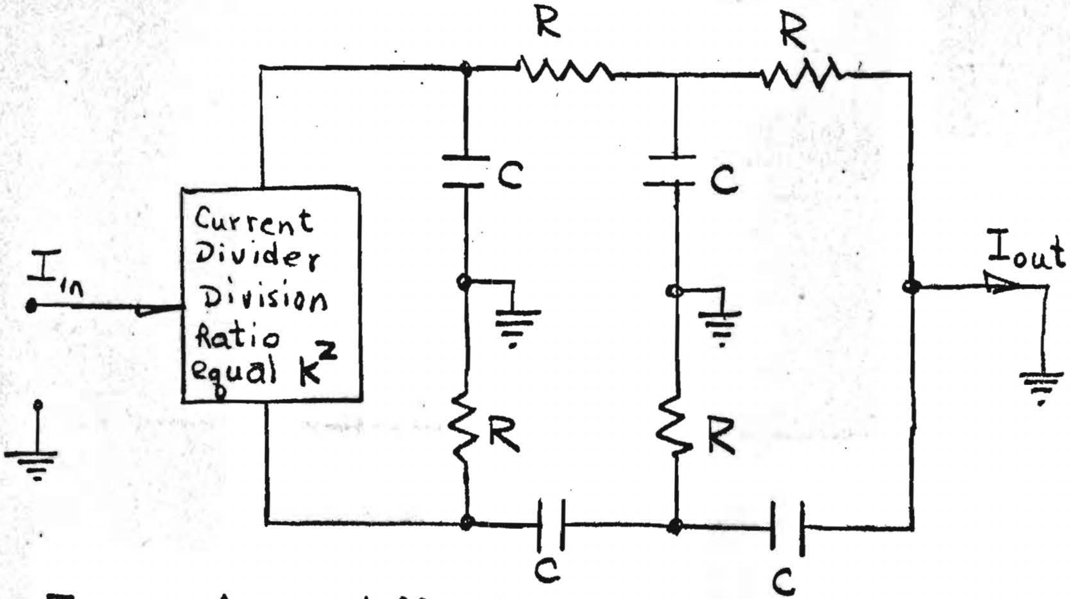


Figure 1 - Null Network

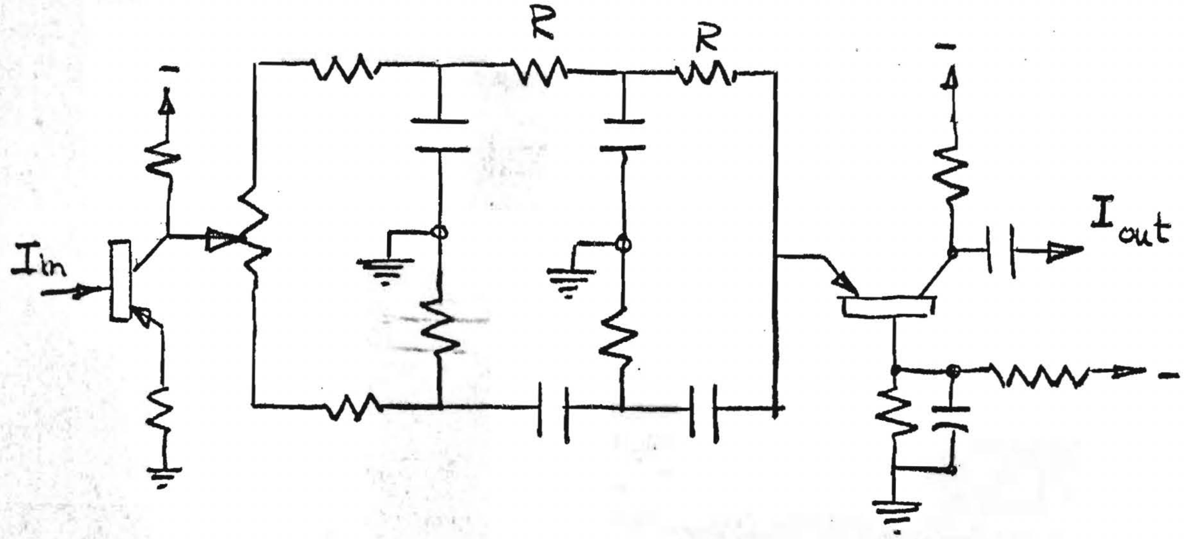


Figure 2 - Practical Null Network

GEORGIA INSTITUTE OF TECHNOLOGY
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

20 September 1961

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

Attention: RCUMA

Subject: Monthly Progress Letter No. 10, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

Project effort during August was directed primarily toward completion of the blanking adaptor in a form suitable for demonstration at RADC in early September.

Several difficulties were encountered in this process. The capacity loaded RF cavity tuner had an unstable tuning mechanism which made it very difficult to set the tuner to the proper frequency. The instability was a result of the piston type capacitor which was used to effect the capacitance loading of the cavity. Apparently this type of capacitor has an inherent mechanical instability in its adjustment mechanism, since several capacitors from various manufacturers were tried, and none was found to be satisfactory.

A new tuner was constructed using a capacitance loaded quarter wave stub; the capacitance in this case being a conventional variable type with air dielectric. The method of construction is illustrated in the sketch of Figure 1. This particular configuration permits a low (50 ohms) input and output impedance to be obtained in an unbalanced manner. However, both sides of the variable capacitor are at an RF potential in respect to ground and, hence, neither side of the capacitor may be grounded. Difficulties from this RF potential were overcome by mounting the capacitor in a polystyrene block and using an insulated shaft to permit tuning. The entire assembly was enclosed in a metal shielded box to isolate the tuning of the circuit from external capacitances.

An AC power supply was constructed to permit the use of the blanking adapter from the 115 volt AC power line, rather than from batteries, as had been done in the laboratory. The necessary voltage regulation for the $\pm$ 6 volt supply was obtained by the use of two Zener diodes placed directly across the output terminal of these two supplies. In addition, a low current $\pm$ 12 volt supply for the operation of the auxiliary receiver was provided. The schematic diagram of the power supply is shown in Figure 2.

Finances:

The amount of funds remaining in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

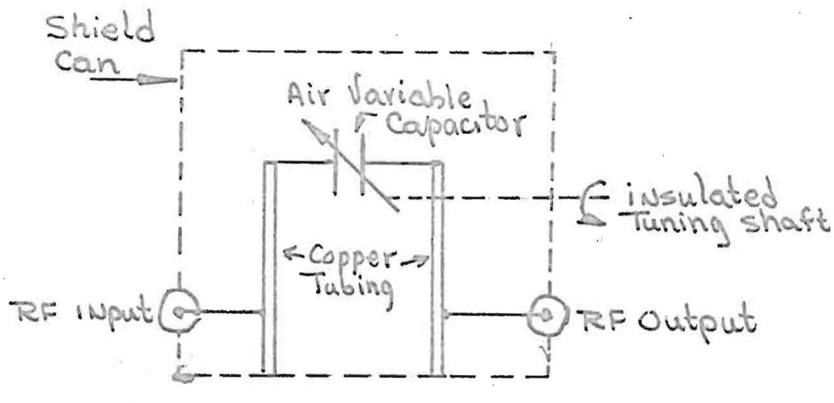


Figure 1 - RF Tuned Filter

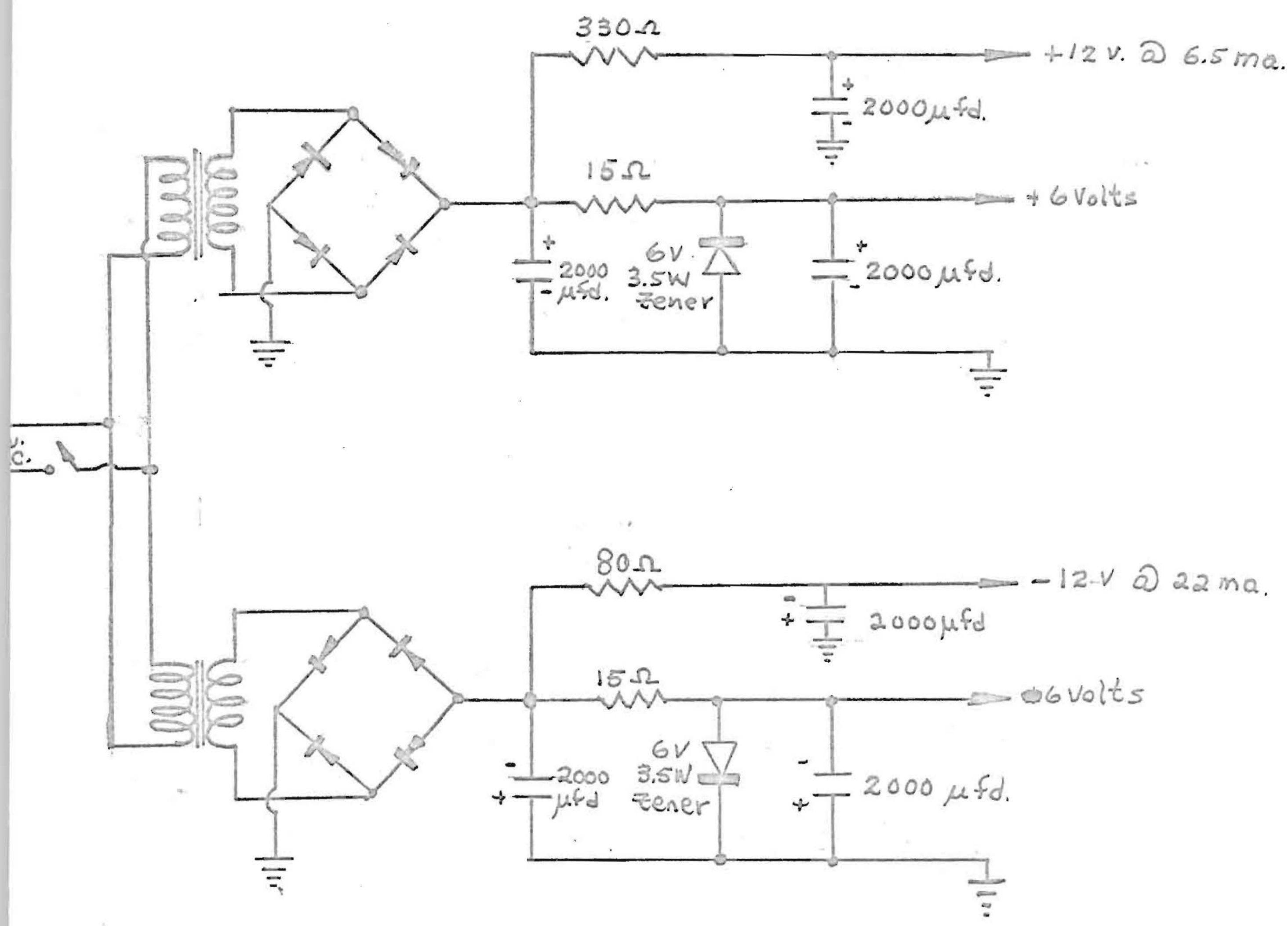


Figure 2 - Power Supply For Blanking Adaptor

GEORGIA INSTITUTE OF TECHNOLOGY

ENGINEERING EXPERIMENT STATION

ATLANTA 13, GEORGIA

18 October 1961

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

ATTENTION: RCUMA

SUBJECT: Monthly Progress Letter No. ~~10~~ 11, Contract AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

Most communications receivers, such as the R361, have AGC detectors which operate on the peak value of the input signal. For the situation where large pulse interference is present, this peak detecting property of the AGC detector will cause an AGC voltage to be developed, which is proportional to the peak of the interfering pulse signal. This results in desensitization of the receiver to the desired signal. The desensitization can be overcome in part by modifying the AGC detector in such a manner that it responds to the average value of the input signal rather than to its peak value.

During September, a scheme was investigated which permits the automatic transfer of the AGC detector from a peak type to an averaging type whenever pulse interference is present. The manner in which this is accomplished is outlined in the Block Diagram of Figure 1. Referring to this figure, the differentiated output of the detector is very small when the pulse interference is absent and as a result the one shot Multivibrator is not triggered and the average value of its output voltage is small. This, in turn, is insufficient to fire the Schmidt trigger circuit and the relay which is controlled by this circuit is in the unenergized position.

As a result, the relay contacts are set so that the AGC circuitry functions in its normal manner. However, when a pulse interference appears at the receiver input, a large output is present to trigger the one shot Multivibrator, causing a change in the average output voltage of this stage, due to the reduced duty cycle in one of the tubes of the one shot Multivibrator. This change in the output voltage fires the Schmidt trigger circuit which, in turn, energizes the relay and transfers the contacts to the other position of the relay.

In this position, the AGC circuit is connected to an averaging type detector, thus developing a much smaller AGC voltage than is normally developed and reducing the desensitization effect of the pulses on the receiver. A schematic of the automatic AGC transfer circuitry developed to date is shown in Figure 2.

Finances:

Funds allocated under Contract No. AF 30 (602)-2366 are almost expended. As of this date, contract extension has not been received.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

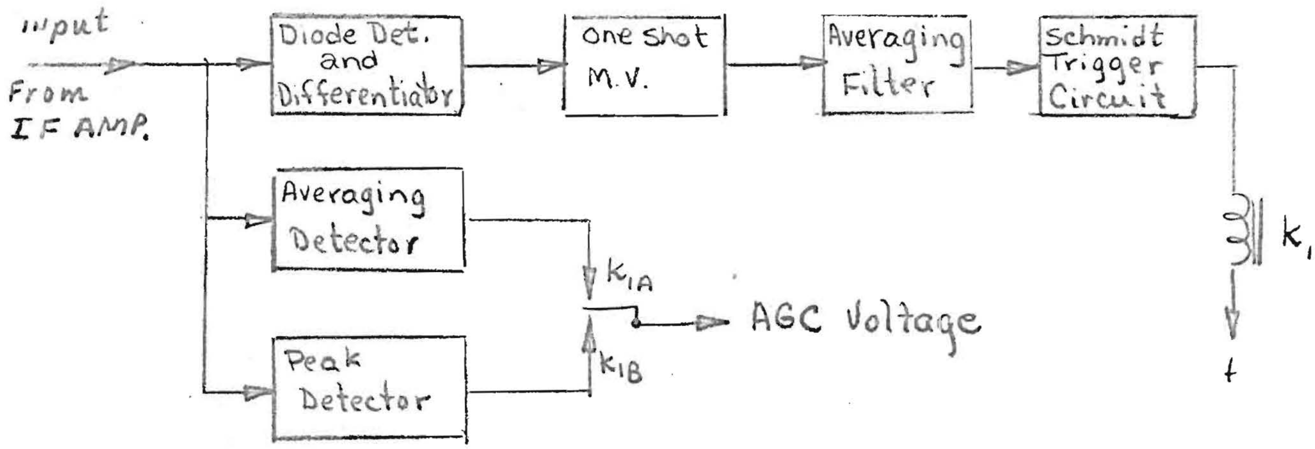


Figure 1. - Block Diagram

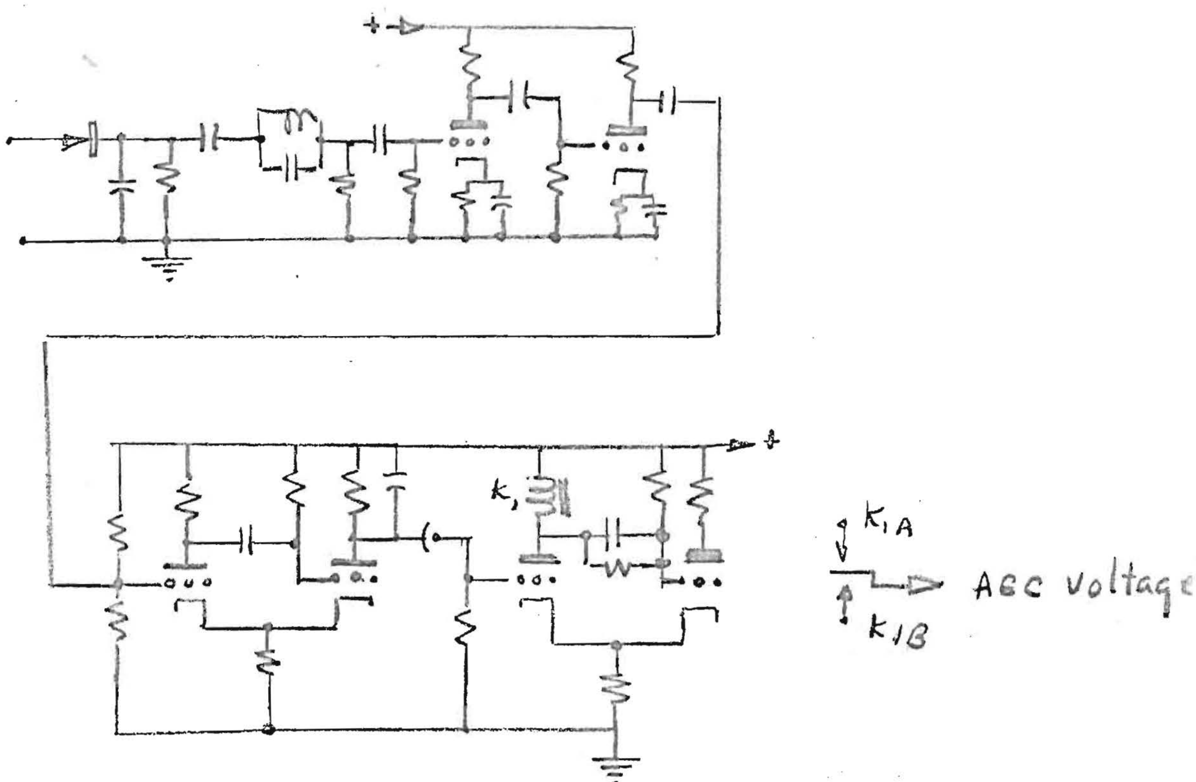


Figure 2. Adaptive AGC Circuitry

GEORGIA INSTITUTE OF TECHNOLOGY
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

21 December 1961

A-525

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

ATTENTION: RCUMA

SUBJECT: Monthly Progress Letter No. 12, Contract AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

Recently there has been considerable attention given to negative resistance devices as low noise amplifiers. One of these devices which we have been investigating, principally as a wide band antenna preamplifier, is the tunnel diode.

There are two main classes of tunnel diode amplifiers which do not employ non-reciprocal elements; the transmission amplifier and the reflection amplifier. Much of the investigation has been directed toward the reflection type since it affords approximately twice the gain bandwidth product of the transmission type and its greater ease in installation.

The transmission type consists of a tunnel diode across the load on the signal source and lossless reciprocal two-port network between the load and signal source. In the reflection amplifier, the signal source and load are directly coupled and are in turn shunted by a lossless reciprocal two-port network that is terminated by the tunnel diode. Figure 1 is a diagram of a reflection amplifier as it might be used in conjunction with an existing antenna and receiver.

The purpose of the passive two-port network is to present the proper negative impedance to the load and signal source to realize the desired power gain over the frequency range of interest. The tunnel diode is not a simple negative resistance but may be represented by the equivalent circuit in Figure 2.

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Where $g_n$ is the absolute value of the negative conductance and Z is the impedance looking into the terminals of the diode,

$$Z = R_s - \frac{g_n}{g_n^2 + (\omega C)^2} + j\omega \left[ L_s - \frac{C}{g_n^2 + (\omega C)^2} \right] .$$

The angular frequency where the real part of the impedance is zero is called the resistive cut-off frequency and is

$$\omega_R = \frac{g_n}{C} \sqrt{\frac{1}{R_s g_n} - 1} .$$

The angular frequency where the imaginary part of the impedance is zero is called the inductive cut-off frequency and is

$$\omega_{I_m} = \frac{g_n}{C} \sqrt{\frac{C}{L_s g_n^2} - 1} .$$

Below the resistive cut-off frequency, the impedance has a negative real part. Above, it has a positive real part. At frequencies above $\omega_R$, there can be no amplification. Below the inductive cut-off frequency, the imaginary part of the impedance is negative, and above it is positive.

For a 1N2939,

$$f_R = 638 \text{ mc},$$

$$\text{and} \quad f_{I_m} = 650 \text{ mc},$$

$$\text{where} \quad f = \frac{\omega}{2\pi} .$$

From this, it is obvious that at low frequencies, the tunnel diode might be approximated by a negative resistor and capacitor in parallel and at high frequencies by a positive resistor and inductor in series.

$$\text{For} \quad R_s = L_s = 0,$$

$$Z = \frac{1}{-g_n + j\omega C} ,$$

which is a negative resistor and a capacitor in parallel. If the upper frequency limit of the low frequency equivalent circuit is defined as the frequency at which

$$10R_s = \frac{g_n}{g_n^{\,2} + (\omega_R{}'C)^2} \quad ,$$

then
$$\omega_R{}' = \frac{g_n}{C}\sqrt{\frac{0.1}{R_s g_n} - 1} \quad ,$$

or
$$10L_s = \frac{C}{g_n^{\,2} + (\omega_{I_n}{}'C)^2} \quad .$$

Then
$$\omega'_{I_m} = \frac{g_n}{C}\sqrt{\frac{0.1C}{L_s g_n} - 1} \quad ,$$

whichever is lower.

For the 1N2939

$$f'_R = 175 \text{ mc.}$$

$$f'_{I_m} = 177 \text{ mc.}$$

Therefore, when operating below $f'_R$ or $f_I{}'_m$, whichever is the lower of the two, the low frequency equivalent circuit is sufficient for most amplifier design calculations except stability analysis. The complete equivalent circuit must be employed when determining the stability of conditions for the circuit.

A necessary and sufficient condition for stability is that the negative resistance be less than the real part of the impedance seen by the negative resistor. Since the tunnel diode is capable of oscillation at frequencies far exceeding the upper limit of the low frequency equivalent circuit, the complete equivalent circuit must be considered in stability calculations.

One simple tunnel diode amplifier is shown in Figure 3. The insertion power gain is defined to be the ratio of power developed in the load resistor with the tunnel diode in the circuit to the power developed in the load resistor without the tunnel diode.

The insertion power gain for the amplifier shown in Figure 3 is

$$P = \frac{(g_s + g_1)^2}{g_t^2 + \omega^2 C^2 \left(1 - \frac{\omega_o^2}{\omega^2}\right)^2} .$$

Maximum power gain is

$$P_{max} = \frac{(g_s + g_1)^2}{g_t^2} ,$$

where

$$g_t = g_s + g_1 - g_n$$

and

$$\omega_o = \frac{1}{LC} .$$

The half power bandwidth of this amplifier is

$$B = \frac{g_t}{C} \text{ in radians/sec.}$$

These calculations have been verified by experiment, and it becomes apparent from the above calculations that extermely large gain bandwidth products are not possible with this simple amplifier. The addition of more tunnel diodes in other circuit configurations to increase the overall gain bandwidth are possible as was found in the literature search. However, with an appropriate two-port network in the reflection amplifier, better overall gain bandwidth products with a single tunnel diode can be realized. The synthesis of these networks require more sophisticated methods and are presently being investigated.

Finances:

The amount of funds remaining in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

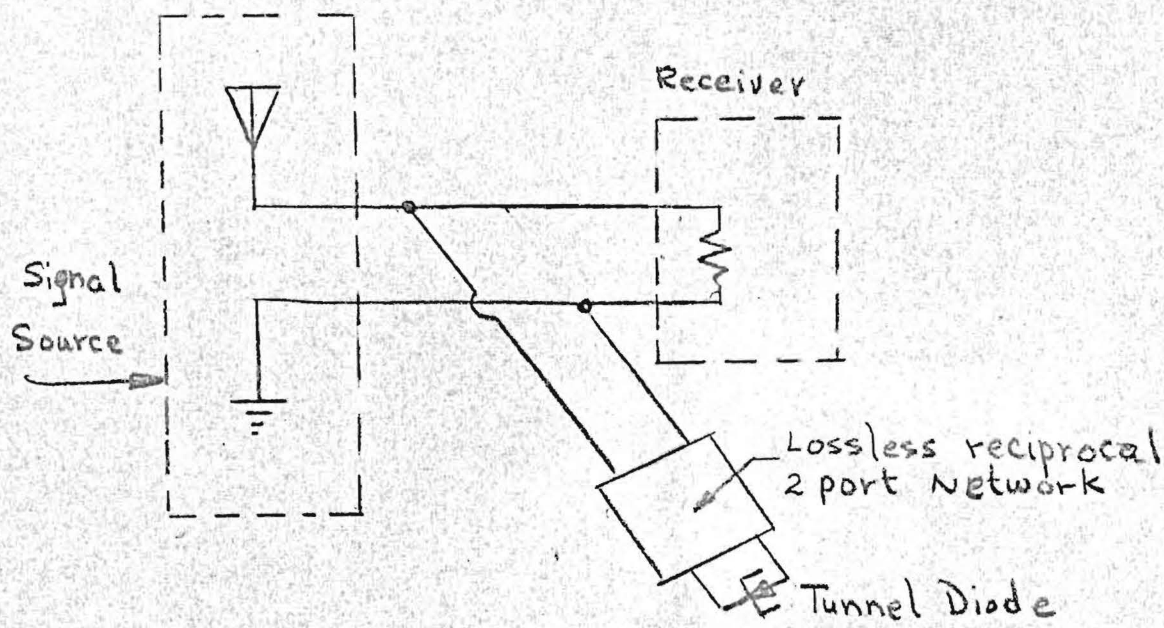



Figure 1. - Reflection Amplifier

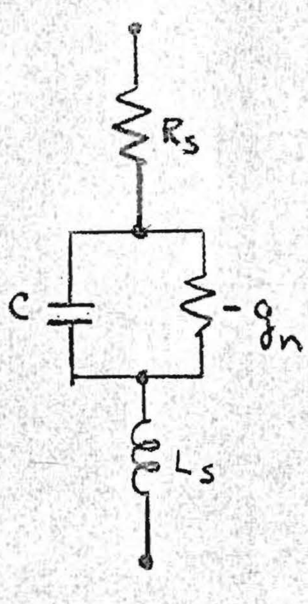
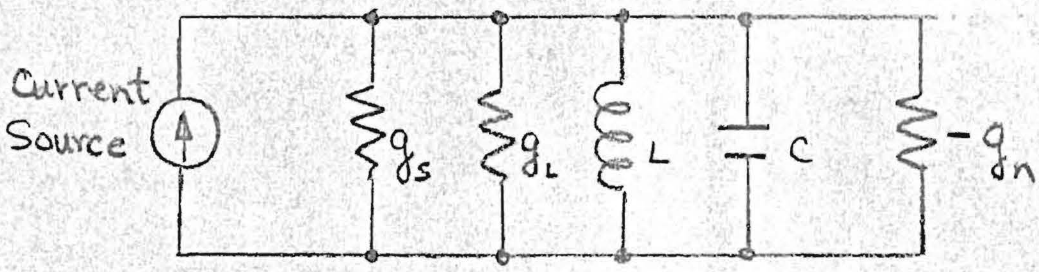



Figure 2 - Tunnel Diode Equivalent Circuit




$g_s$ = Source Admittance
$g_L$ = Load Admittance

Figure 3 - Tunnel Diode Amplifier

A-525
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**GEORGIA INSTITUTE OF TECHNOLOGY**
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

22 January 1962

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

Attention: RCUMA

Subject: Monthly Progress Letter No. 13, Contract AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

A well known relationship used to determine the possible responses of a frequency converter is the one shown in Equation (1):

$$\omega_S = \frac{\omega_{IF} \pm P\omega_{LO}}{Q} . \qquad (1)$$

Where: $\omega_S$ is the frequency at which the response can occur,

$\omega_{IF}$ is the intermediate frequency of the receiver,

$\omega_{LO}$ is the local oscillator frequency, and

P and Q are integers.



In an ideal mixer, P and Q have only the value unity which corresponds to the desired and image responses. All other values of P and Q correspond to spurious responses; the P values being associated with harmonics of the input signal, while the Q's are associated with harmonics of the local oscillator signal. There are two general methods by which the effects of these spurious responses can be reduced. One method is to place adequate frequency selectivity in the signal input to the mixer, so that only the desired signal can reach the input terminals of the mixer. The spurious responses are still present but since no signals appear in the mixer input terminals at the spurious frequencies, no spurious outputs can result.

A second method is to eliminate as many values of P and Q as possible. During December, some attention was given to reducing spurious responses in a mixer by the elimination of as many values of Q as possible. The sketch in Figure 1 illustrates one scheme for accomplishing the desired result. The transmission of a desired signal is alternately switched by the local oscillator from one or the other of two linear states. Since both of the states are linear, no harmonics of the input signal can be generated and, consequently, there are no values for Q above one. This is strictly true only if the time of switching from one state to the other is zero. In a practical realization of such a mixer, there is a finite switching time involved and a consequent generation of harmonics of the input signal. However, the amplitudes of these harmonics are considerably lower than are normally encountered in a conventional mixer.

Figure 2 shows the practical realization of the switching mixer which was used to determine the performance which could be obtained. Initial tests indicate that (at least at the lower frequencies at which these tests were made) a reduction of 20 to 30 db in the level of spurious responses can be obtained over conventional type mixers. Further study during January should permit a more accurate determination of the relative advantages of such a mixer.

As mentioned in our last letter, a simple tunnel diode amplifier has a very limited gain bandwidth product and more complex circuitry is required to obtain sufficiently large gain bandwidth product. As a result of this increased complexity, more sophisticated methods of analysis and network synthesis are required. In conjunction with a study of advanced design techniques, an intensive literature search has been carried out in surveying the state of the art in tunnel diode amplifier design. Gain bandwidth products using non-reciprocal elements or curcuit configurations employing more than one diode can yield larger gain bandwidth products than the single diode type. However, the reflection type amplifier shown in Figure 3 has a power gain bandwidth in excess of 6,000. The equalizer design values shown in Figure 3 with a tunnel diode of $g_n = 40 \times 10^{-3}$ mhos and C = 26 pf will yield a gain of approximately 15 db from 180 mc down to the lower frequency limit of the transformer.* Tunnel diodes of the above specifications have been ordered and the amplifier will be fabricated when the diodes are received.

Finances:

Assuming receipt of contract extension, the amount of funds in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

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* Smilen, L.I. and Youla, D.C., "Exact Theory and Synthesis of a Class of Tunnel Diode Amplifier", Proceedings of the National Electronic Conference, Oct. 1960, Vol. XVI.

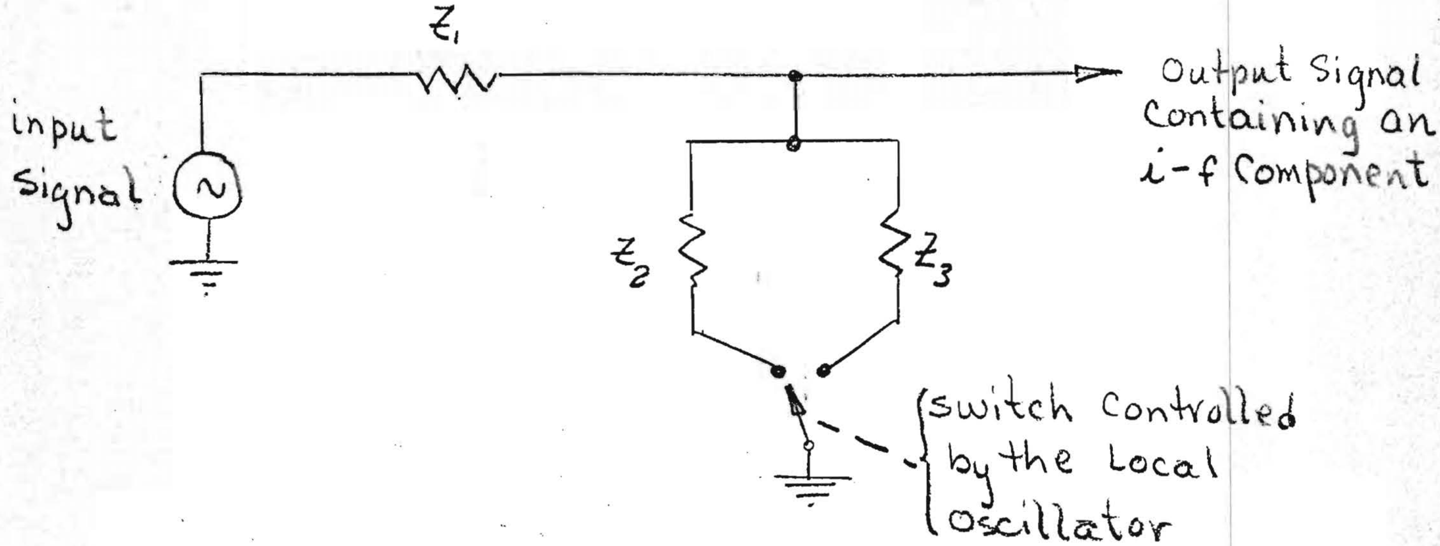



Figure 1 - Switching Mixer

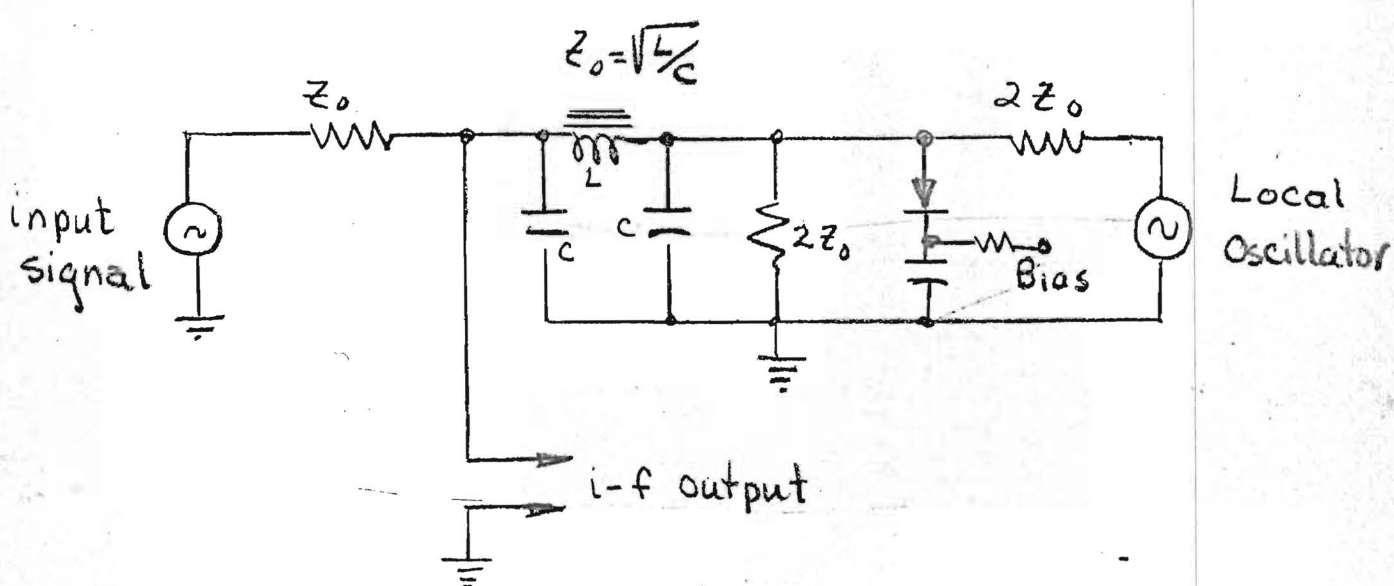



Figure 2 - Actual Mixer Circuit

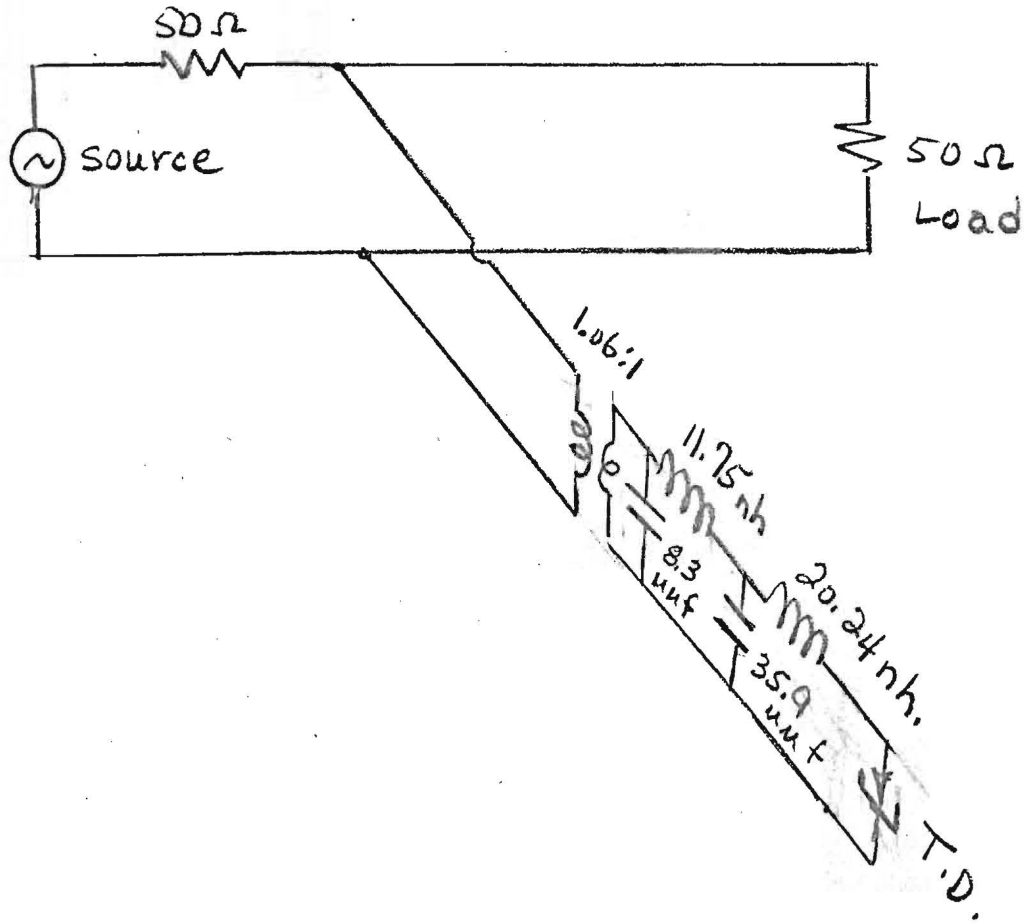



Figure 3 - Reflection Amplifier

GEORGIA INSTITUTE OF TECHNOLOGY
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

23 February 1962

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

Attention: RCUMA

Subject: Monthly Progress Letter No. 14, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

In January, efforts were concentrated on the improvement of spurious response rejection in mixers. Tests made on the previously mentioned switching mixers demonstrated that the magnitude of the responses due to the generation of harmonics of the input signal are directly related to the switching time. As expected, the faster switching times were associated with the smaller harmonic production, since this rapid switching minimizes the fractional switching time in which the mixer is in a nonlinear condition.

In addition, a technique was devised which should permit reduction in the amplitude of the spurious responses due to the generation of harmonics of the local oscillator signal. This technique is basically an extension of the cancellation arrangement found in the conventional balanced mixer. In a balanced mixer, the responses due to the even harmonics of the local oscillator are cancelled at the IF frequency because the local oscillator signals fed to the two mixers are 180 degrees out of phase. However, no cancellation of the odd harmonics responses is possible with such an arrangement.

If a pair of balanced mixers, i.e. a total of four mixers, are driven with local oscillator signals which are in quadature (namely the local oscillator signals to the mixers are at phase angles of 90, 180, and 270 degrees), then by proper combination of the IF outputs of this pair of balanced mixers, it is possible to cancel all spurious responses, both even and odd order, up to order five. More complicated arrangement of mixers offers the possibility of cancellation of spurious responses through orders higher than five. However, the practical problem of maintaining the proper amplitude and phase angles for cancellation using a large number of mixers would, in all probability, preclude practical application of this technique.

The mixing system for cancellation of spurious responses to order five is shown in Figure 1. The method by which the outputs of the four mixers are combined to produce the desired cancellation can be seen by an inspection of the vector diagrams of Figure 2.

Finances:

Assuming receipt of contract extension, the amount of funds in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

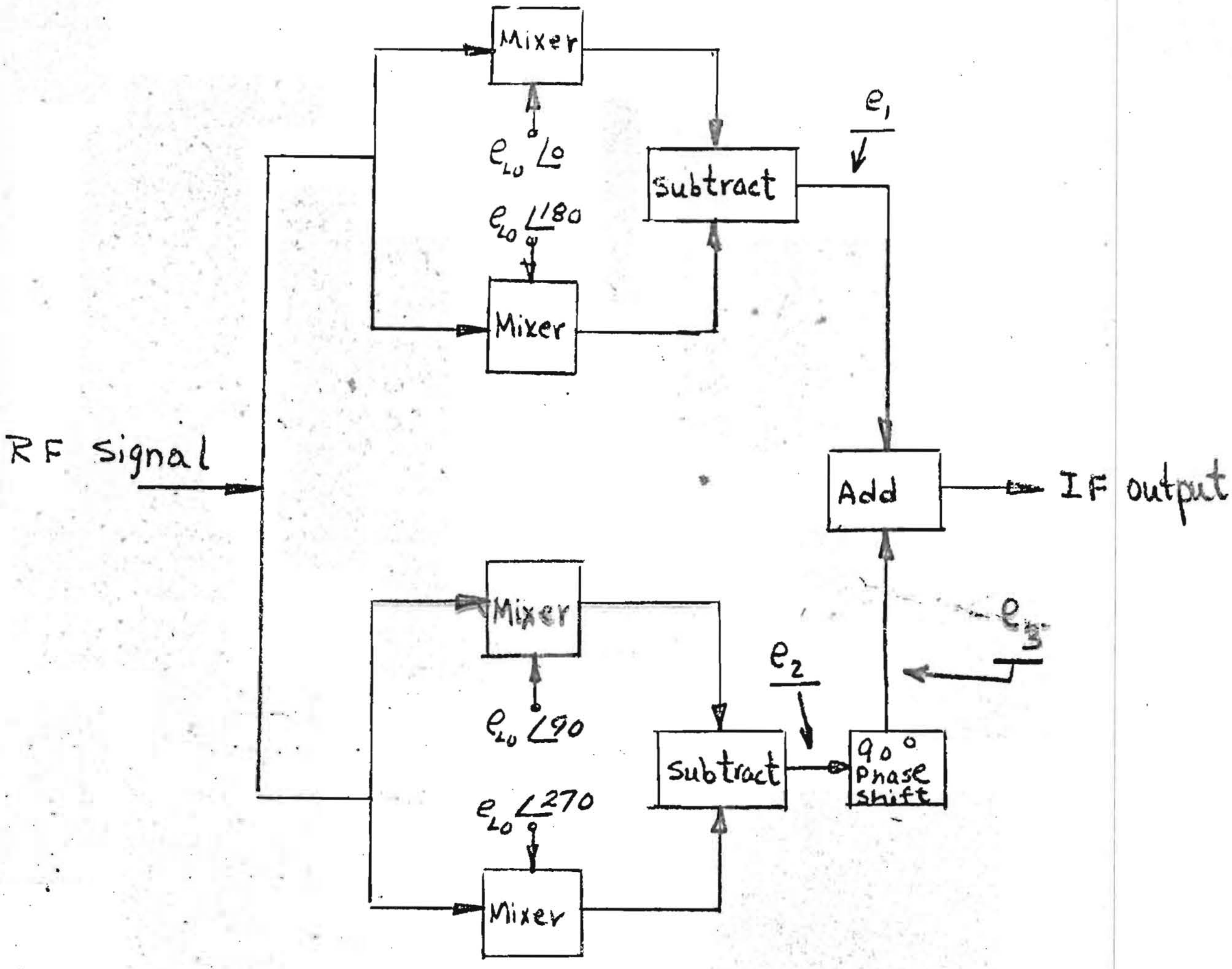



Figure 1- Cancellation Mixer

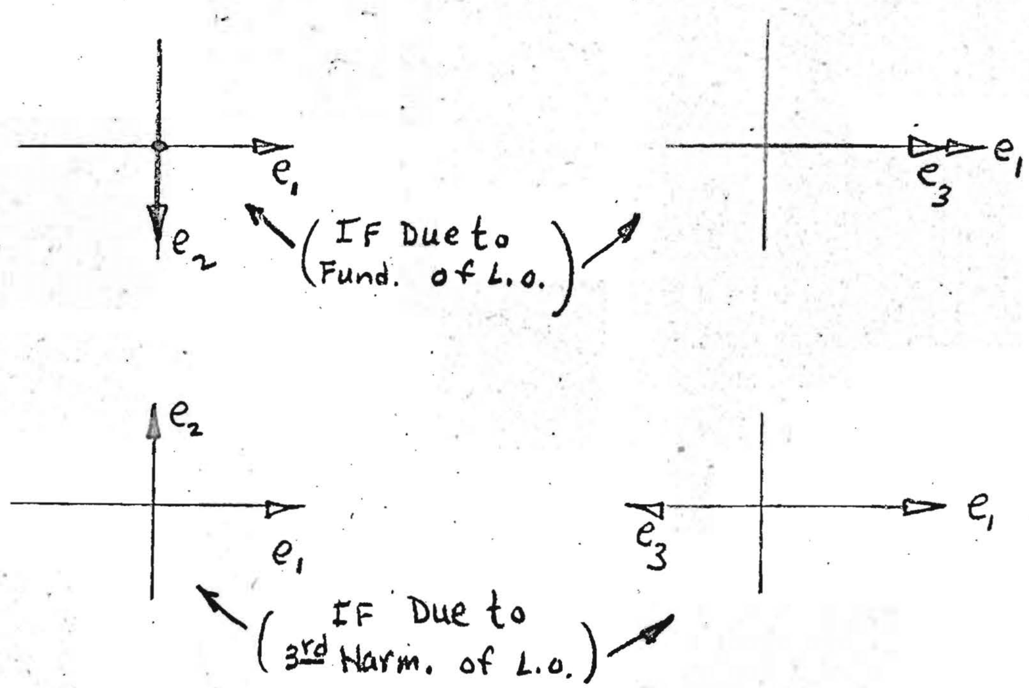



Figure 2- Vector Diagrams showing Cancellation

GEORGIA INSTITUTE OF TECHNOLOGY
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

23 March 1962

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

Attention: RCUMA

Subject: Monthly Progress Letter No. 15, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

During February, several mixer circuits were constructed to determine their relative ability to cancel even-order spurious responses. The simple mixer circuit of Figure 1 exhibited a cancellation of 60 db on the second harmonic local oscillator response and, to a lesser degree, cancellation of spurious responses due to higher even-order harmonics of the local oscillator signal. Some of the additional advantages to be gained by the use of such a balanced mixer can be seen from the familiar equation relating the frequency of a spurious response to the local oscillator and IF frequencies.

$$\omega_S = \frac{P\omega_{LO} \pm \omega_{IF}}{Q} \qquad (1)$$

The desired response occurs for P = Q = 1 so that:

$$\omega_D = \omega_{LO} - \omega_{IF} \qquad (2)$$

The frequency separation between the desired response and the spurious responses then is:

$$\Delta\omega = \omega_D - \omega_S = (\omega_{LO} - \omega_{IF}) - \frac{(P\omega_{LO} \pm \omega_{IF})}{Q} \qquad (3)$$

or:

$$\Delta\omega = \omega_{LO}(1 - \frac{P}{Q}) - \omega_{IF}(1 \pm \frac{1}{Q}) \qquad (4)$$

If P = Q, the first term of Equation 4 is zero and if, in addition, Q = 2, then:

$$\Delta\omega = \frac{\omega_{IF}}{2} \qquad (5)$$

This response is very close to the desired frequency and, for a small value of $\omega_{IF}$, is difficult to eliminate with selectivity in front of the mixer. However, a balanced mixer will cancel this response because it is of even order.

Notice that, for P = Q, at higher values of Q, the value of $\Delta\omega$ becomes:

$$\Delta\omega_{\substack{P = Q \\ Q \to \text{Large}}} \approx \omega_{IF} \qquad (6)$$

This again points out the value of using a high IF frequency.

Work has begun on the construction of a digital delay line for audio signals due to the difficulty of obtaining a satisfactory passive delay line. The delay will be obtained by pulse position modulating a 12 kc pulse carrier and delaying this modulated pulse train with a series of one-shot multivibrators. The demodulated output of these multivibrators is a delayed audio signal. The block diagram of Figure 2 illustrates this technique.

Finances:

The amount of funds in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

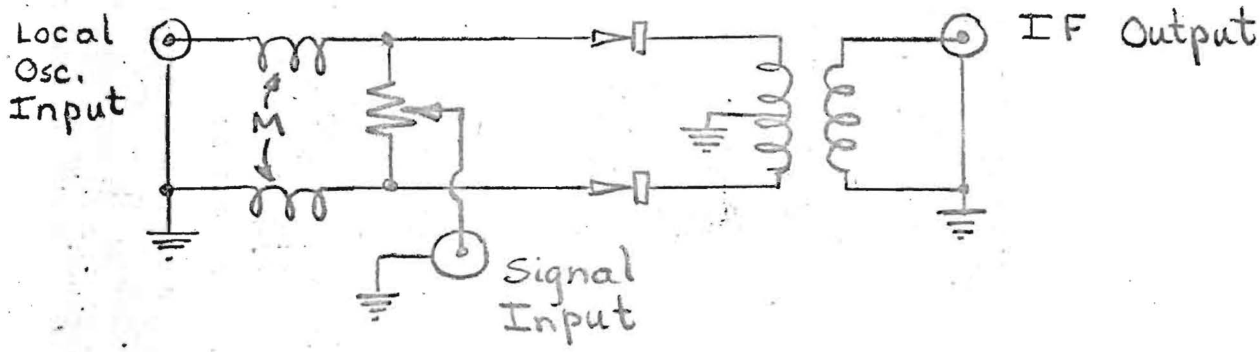



Figure 1- Balanced Mixer

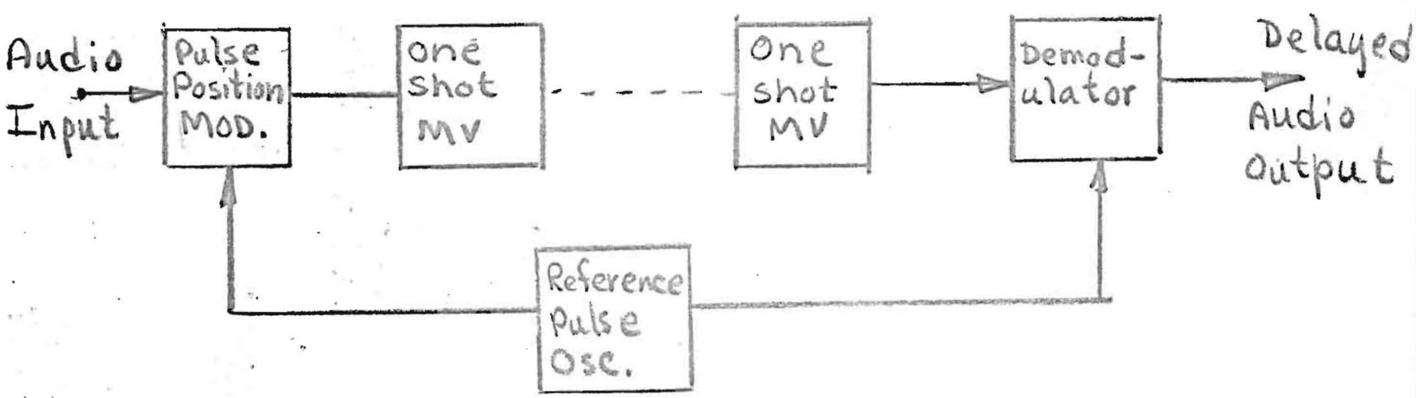



Figure 2 - Digital Delay Line

# GEORGIA INSTITUTE OF TECHNOLOGY

ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

17 April 1962

A 525

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

Attention: RCUMA

Subject: Monthly Progress Letter No. 16, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

Work was continued in March on the design of mixer circuits. The suppression of spurious responses of three mixers was investigated, and the results obtained are summarized in the attached curves. The circuits of the three mixers are shown in Figure 1.

Mixer No. 1 is a single diode type without balancing and is included to serve as a reference with which the other two mixers may be compared. Mixer No. 2 is the one reported in the progress letter for February and is balanced for even harmonics of the local oscillator signal. Mixer No. 3 is a four diode mixer which is balanced for even harmonics of both the signal and the local oscillator.

Referring to the curves, the desired response is fixed at 50 mc, and its level is taken as a 0 db reference. The absolute value of this level is -61 dbm. The local oscillator is fixed at 80 mc. The levels of the spurious responses are the relative levels, with respect to the desired signal, which are required to produce a response equal to that of the desired signal.

The curves show that in every case the single diode mixer is inferior to the two balanced types. Of the two balanced types, the doubly balanced mixer is consistently better than the singly balanced one. It is concluded that improvement of the order of 30 db can be obtained by the use of careful balancing; however, it is felt that the low conversion gain associated with a diode mixer will require that a vacuum tube mixer be used in any final equipment which is constructed. Nevertheless, the same balancing techniques are applicable to vacuum tube mixer circuitry and the same improvements in spurious response rejection should be obtained. All of the above data were taken with no selectivity in front of the mixer. The addition of such selectivity can be expected to increase the rejection of spurious responses drastically.

The design of the circuitry for the digital delay line mentioned in last month's letter is approximately 50 per cent complete. A typical delay section is the one-shot multivibrator shown in Figure 2. The delay of this particular section is adjustable from 50 to 70 microseconds. The delay per section is controlled by the rate at which samples of the input signal are taken. For our purposes, we have chosen a sampling rate of approximately 12 kc. This permits an audio bandwidth of 5 kc to be sampled at slightly greater than the Nyquist rate of 10 kc with sufficient guard band to permit simple filtering at the output of the delay line. Since the period of the sampling rate is 83.3 microseconds and since the recovery time of the multivibrator is a minimum of 10 per cent of the delay, the maximum delay obtainable is approximately 70 microseconds.

To obtain a total delay of 1,000 microseconds will require approximately 15 sections of this delay. The sections will be built on separate plug-in cards. A demodulator for the output of the pulse delay section is shown in the circuit of Figure 3. This demodulator operates as follows: A linear sawtooth voltage is initiated by the reference pulse and is sampled at a later time by the delayed pulse. The value of this sample is stored on the capacitor C. Hence, the voltage stored on the capacitor is a linear function of the time of the delayed pulse with respect to the reference pulse. The voltage across the capacitor is low pass filtered to recover the delayed signal.

Finances:

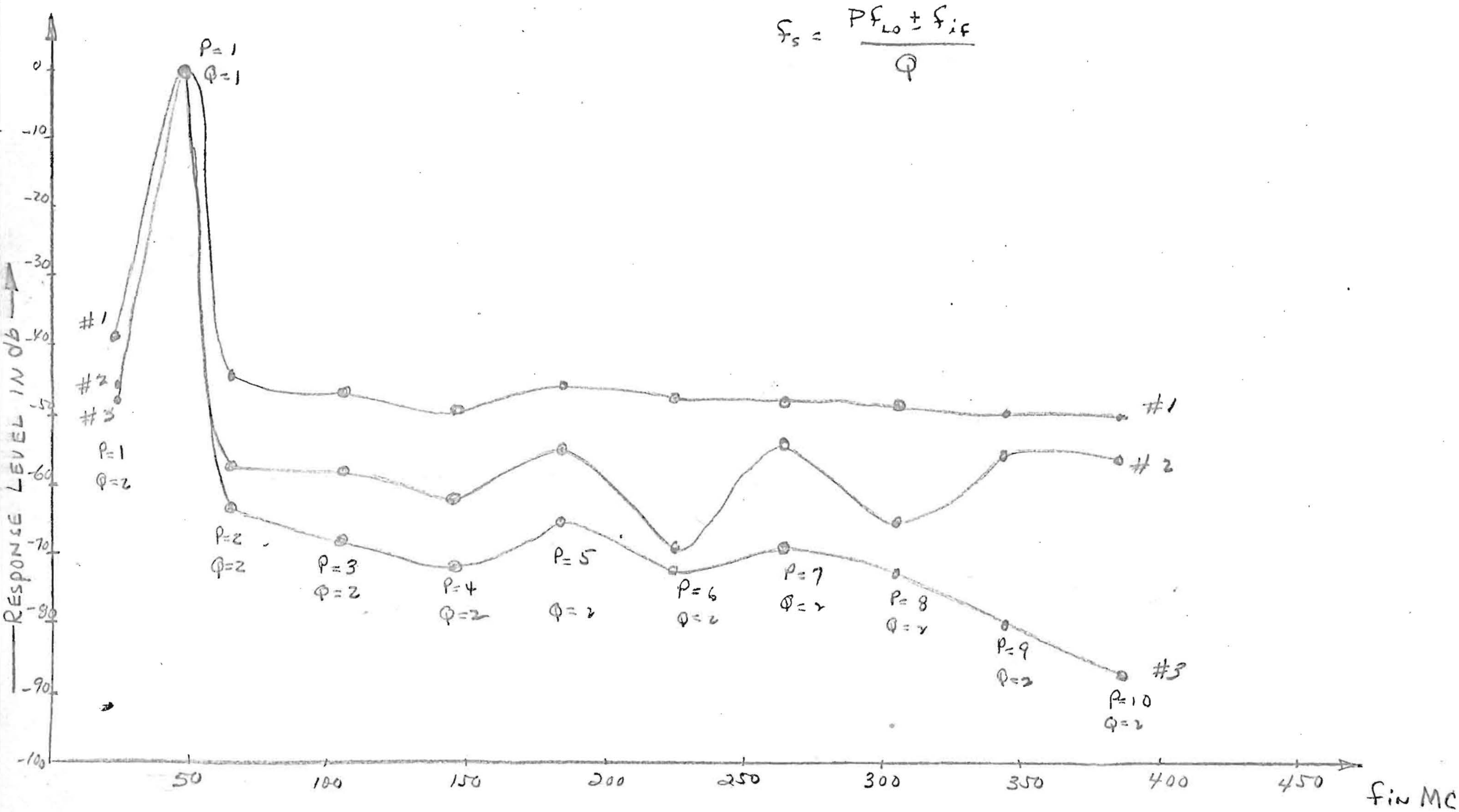
The amount of funds in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

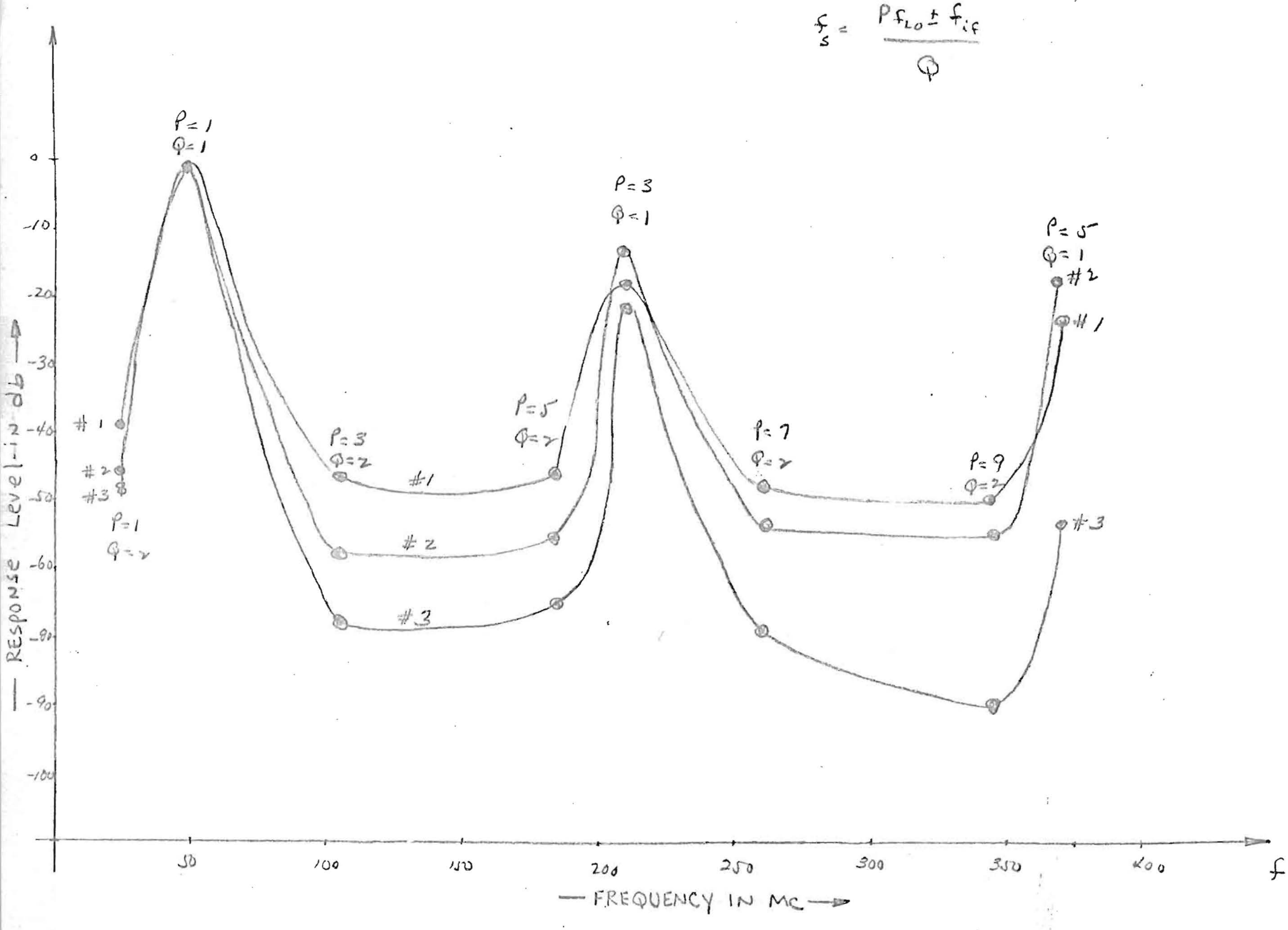
W. B. Warren, Jr.
Project Director

Approved:

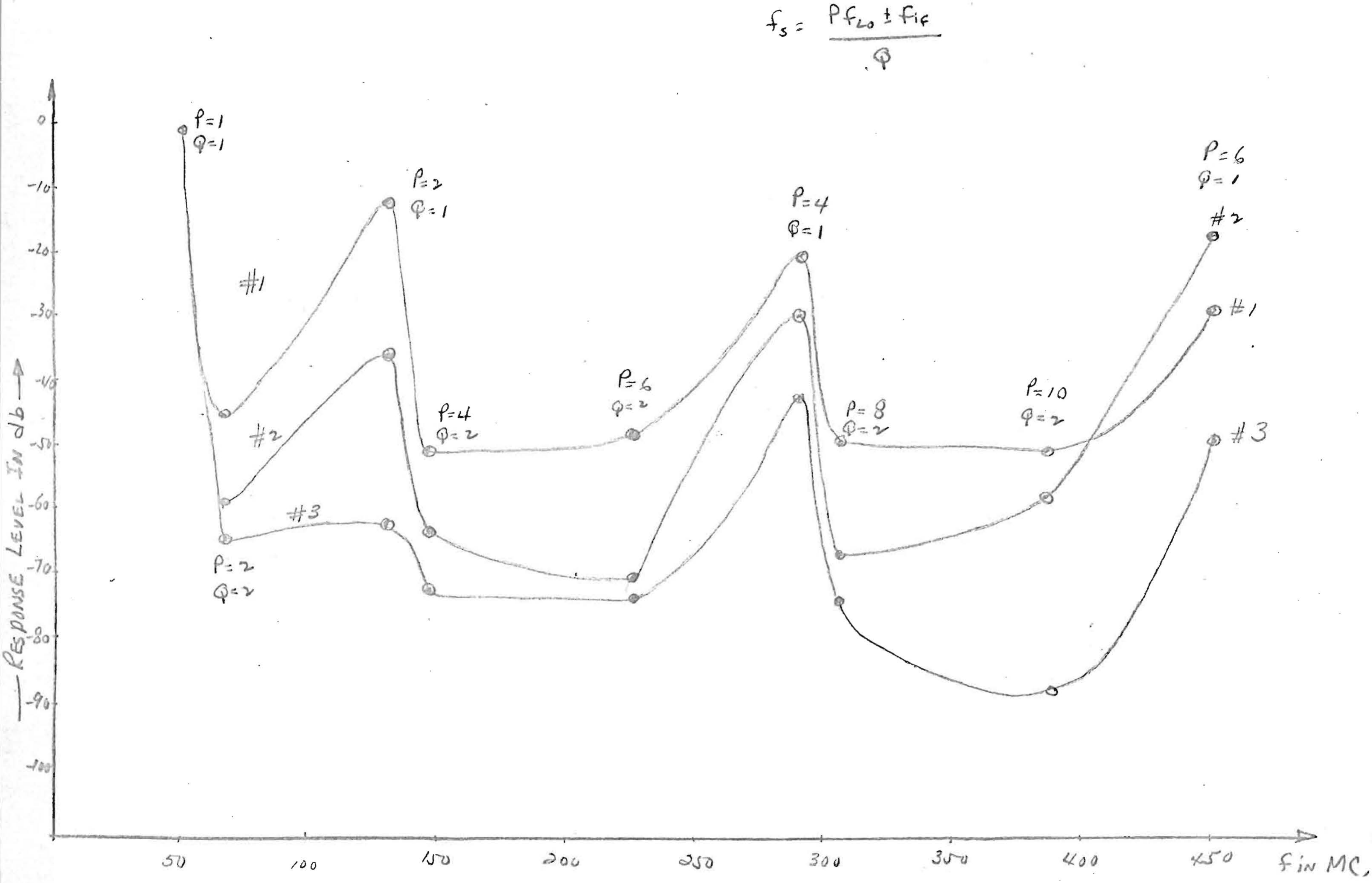
for W. B. Wrigley, Head
Communications Branch



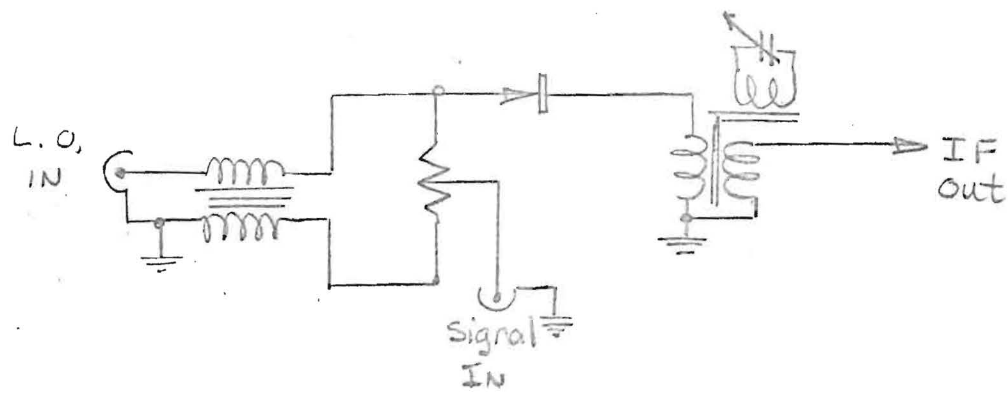

MIXER SPURIOUS RESPONSE LEVELS FOR Q = 2



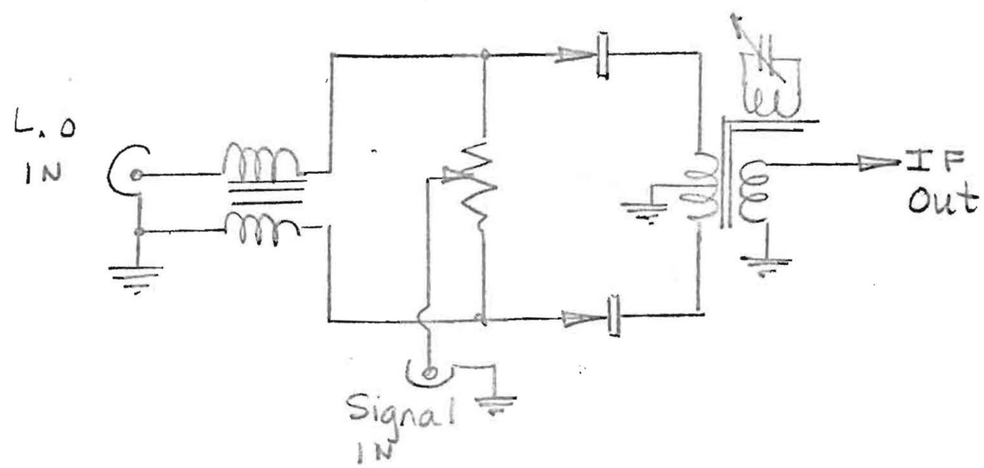

MIXER SPURIOUS RESPONSE LEVELS FOR ODD VALUES OF P




MIXER SPURIOUS RESPONSE LEVEL FOR ...




Mixer # 1




Mixer # 2

Mixer #3

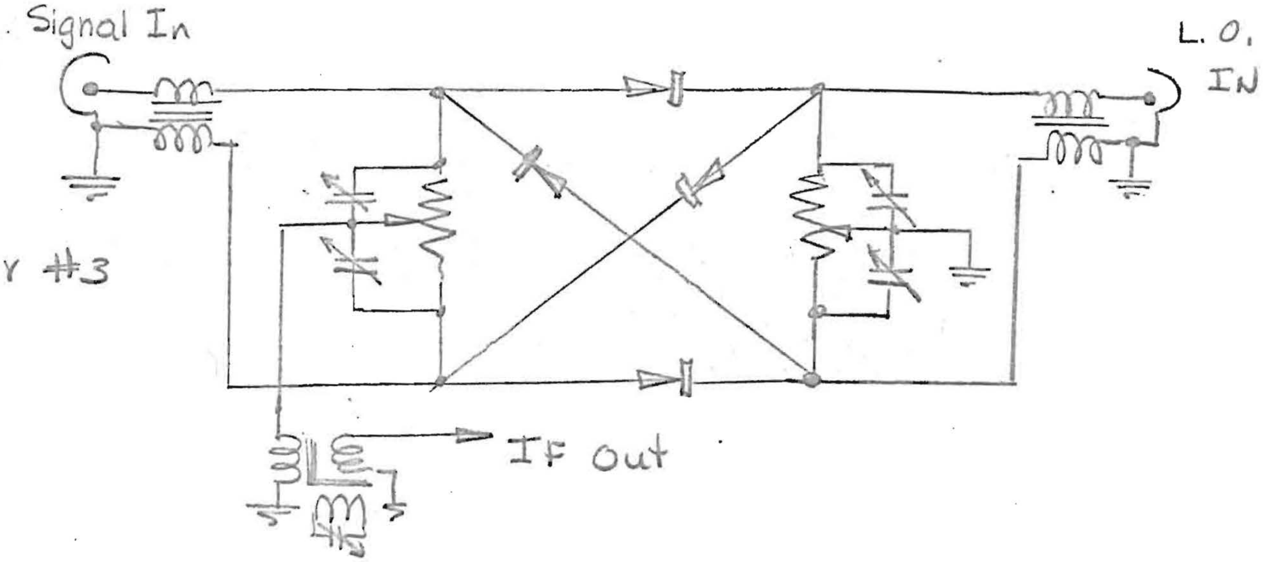



Figure 1 – Mixer Circuits

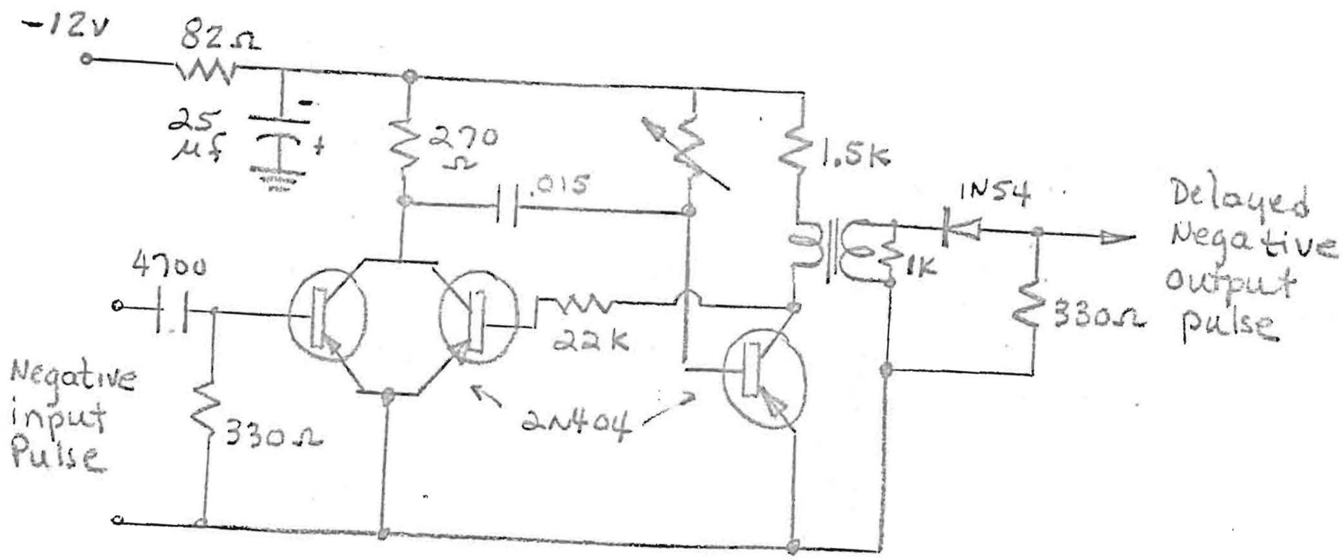



Figure 2 - Multivibrator Delay section

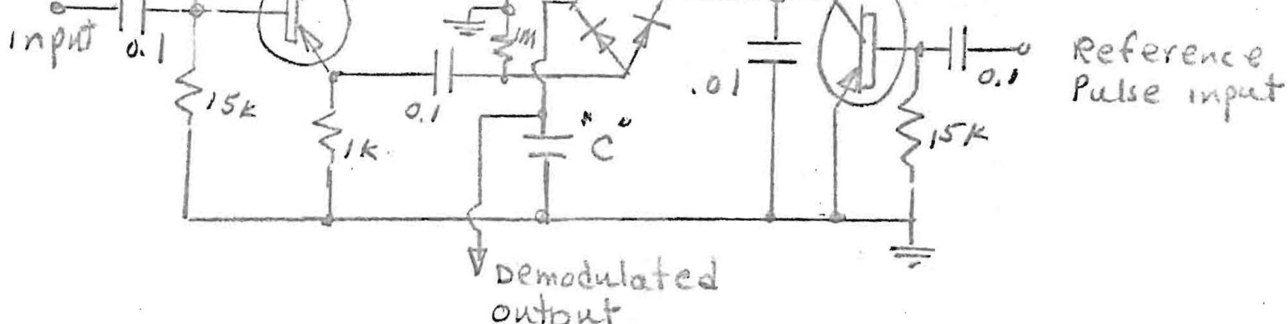



Figure 3 - Pulse Position Demodulator

GEORGIA INSTITUTE OF TECHNOLOGY

ENGINEERING EXPERIMENT STATION

ATLANTA 13, GEORGIA

18 May 1962

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

A 525

Attention: RCUMA

Subject: Monthly Progress Letter No. 17, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

The attached Figure 1 illustrates a cavity preselection filter which was constructed during April. It is planned to use this filter as an input preselector to improve the spurious response rejection of the equipment being constructed for use with field intensity meters.

The tuning range of the preselector is approximately 200 to 500 mc, which is more than adequate to cover the 200 to 400 mc tuning range of the NF105 field intensity meter. Referring to Figure 1, the conical top loading arrangement placed at the end of the center conductor has been provided to lower the rate at which capacitance is added as a function of center conductor length. Specifically, the addition of a simple disc-shaped "top hat" would provide too rapid a change in capacitance at the lower end of the tuning range so that the tuning rate at the low frequency end would be very fast. The use of the cone-shaped "top hat" permits the change in capacitance to be more gradual, with a consequent reduction in the tuning rate. Simple link coupling is used between two of these cavities to provide the selectivity curve shown in Figure 1. Careful machining of the two cavities makes possible the accurate tracking of one cavity with the other.

A grounded-grid RF amplifier and balanced mixer have been constructed during April. The sensitivity of this combination is approximately one-half microvolt and when used in conjunction with the preselector previously described, the strongest spurious response noted was at a level of 95 db above the response due to the desired signal.

Some difficulty has been encountered in constructing a satisfactory local oscillator for use with the balanced mixer. Several oscillators have been constructed which perform satisfactorily at fixed frequencies but all have points in their tuning range at which oscillation stops so that they are unsatisfactory for our use. In addition, it has been determined that a carefully filtered oscillator signal will be necessary to permit the full balancing action of the mixer on harmonic responses to be obtained. This is due to the fact that even harmonics of the local oscillator which are contained in the local oscillator signal cannot be balanced out in the mixer since they are not internally generated in the mixer itself. Tests show that an improvement in suppression of responses due to mixing with even harmonics of the local oscillator can be reduced by 10 to 20 db by careful filtering of the local oscillator output signal before it is applied to the balanced mixer. Such filtering is not generally necessary in a conventional single-ended mixer, since the level of internally generated harmonics normally exceeds those supplied by the local oscillator by many decibels.

Work continued on the digital delay line for use in the audio filter. In the past month, improvements have been made in the linearity of the pulse position modulator and demodulator and several variations of the basic one-shot multivibrator delay section have been tried in an effort to minimize the drain of this device on the power supply. It is felt that this is important since small savings in power at each unit will be multiplied by the number of units used. Present plans are to include approximately 25 of the basic one-shot delay sections.

A chassis and mounting box has been purchased to permit the construction of the entire audio filter on a plug-in, printed circuit card basis. This will permit the simple addition of more delay sections if it is felt necessary.

During the month, two quartz delay lines and an additional R361 receiver were received from RADC. Initial tests indicate that the delay lines have satisfactory phase response to permit their use in an audio cancellation filter. The insertion loss associated with these delay lines is quite large, being in the neighborhood of 50 to 60 db. This should not pose a serious problem, however, since equal attenuation of the undelayed signal by this amount is easily obtained. Amplification of the two channels after they are combined will reduce the signal level to its original value. Tests are contemplated to compare the effects of pulse interference on the previously modified R361 receiver with the unmodified one just received.

It is felt that these tests will bear out the contention that the modifications already reported are effective in improvement of the suppression of pulse interference in these receivers.

Finances:

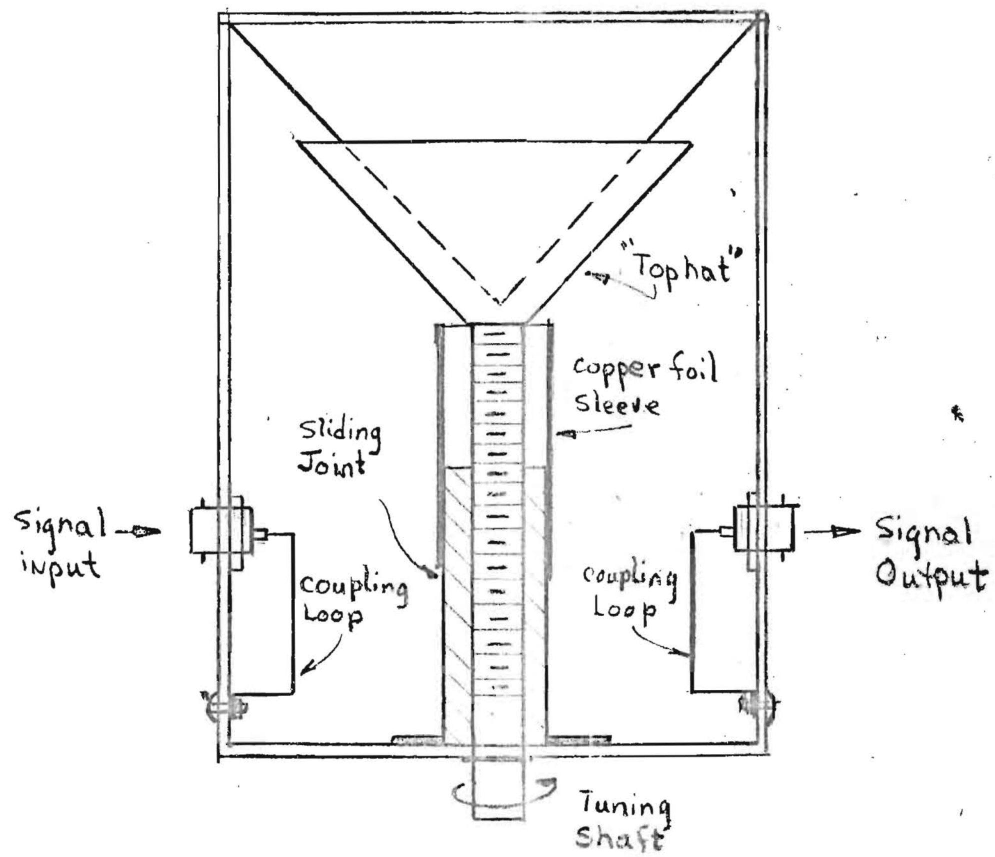
The amount of funds in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch



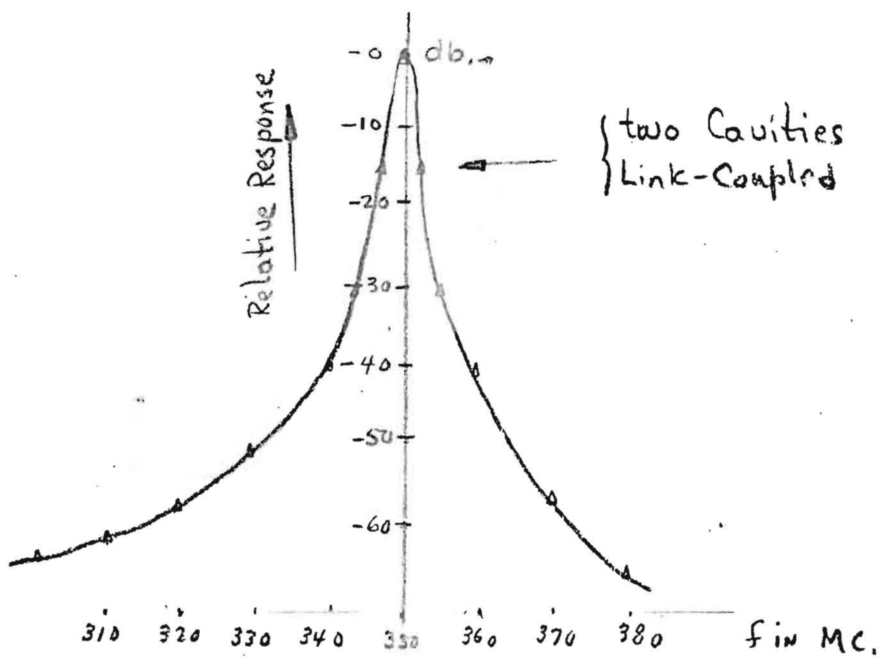




Figure 1 - Cavity Preselector

A525

GEORGIA INSTITUTE OF TECHNOLOGY

ENGINEERING EXPERIMENT STATION

ATLANTA 13, GEORGIA

12 June 1962

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

Attention: RCUMA

Subject: Monthly Progress Letter No. 18, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

Many of the problems mentioned in the April progress letter concerning the construction of a satisfactory local oscillator system have been overcome. The form in which this local oscillator has been constructed can be seen in Figure 1.

Since two outputs of equal amplitudes and opposite phase are required to drive the balanced mixer, the oscillator has been constructed in a push-pull configuration. This permits the use of an inherently balanced system. To insure that only those harmonics which are generated in the mixer itself are effective in producing spurious responses, a filter has been provided at the local oscillator output. This filter is slot coupled to the oscillator output through a pair of slots in the divider wall between them. Two slots are used in order to maintain the balance of the system. The two output coupling links are attached to the sliding shorting bar which tunes the filter to the frequency of the oscillator. The shorting bars for both the filter and the oscillator are positioned by means of two threaded rods which, in turn, are coupled by gears to a common idler gear. This permits one control to tune the oscillator and the filter simultaneously. Tests show that the harmonic suppression as well as the phase and amplitude balance of the outputs of this oscillator is adequate for our use.

A final circuit for the typical delay section for use in the audio filter has been determined. The circuit for this delay section is shown in Figure 2. This delay section has been laid out for a printed circuit type wiring, and ten plug-in cards containing three delay sections each, are in the process of construction.

Initial plans are to test the delay by a simple subtraction network as shown in Figure 3. However, it is felt that less distortion of the desired signal can be obtained by the modification shown in Figure 4. In this arrangement, the signal rather than the interference would be eliminated. The incorporation of controlled positive feedback in the delay path will permit excellent rejection of the desired signal so as to provide an output which is essentially the interfering signal. If this output is now subtracted from the total signal plus interference, the result is the desired signal.

Finances:

The amount of funds in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

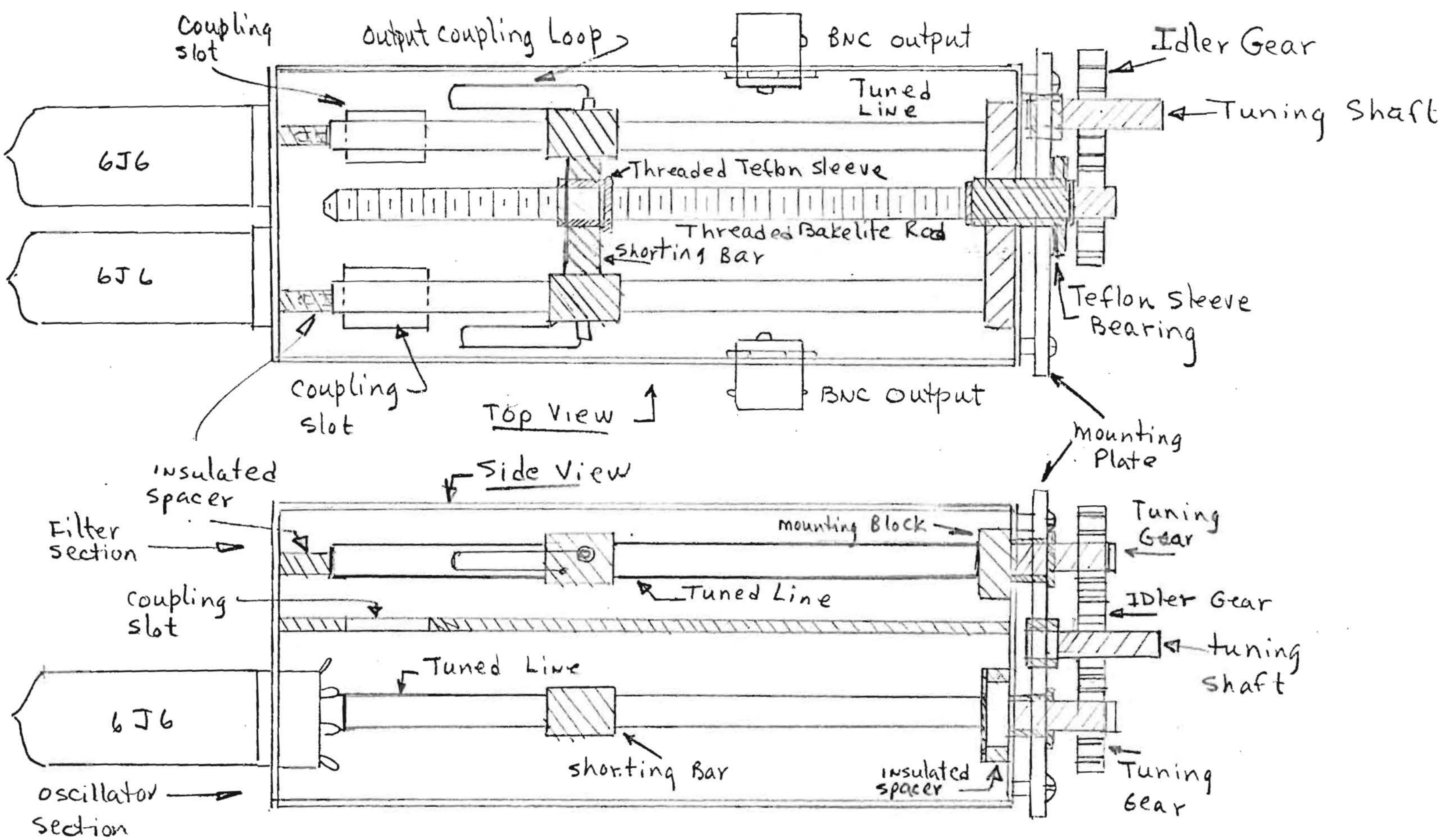



Figure 1 - Local Oscillator Assembly

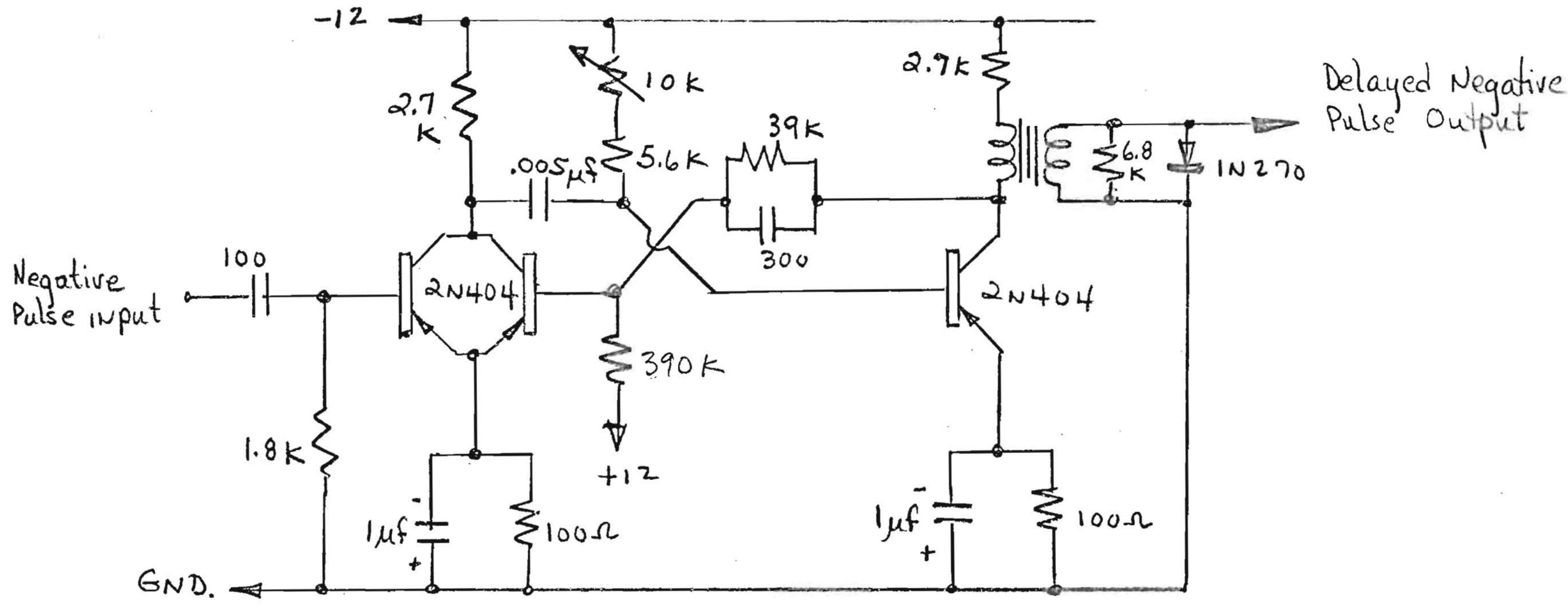



Figure 2 - Delay Multivibrator

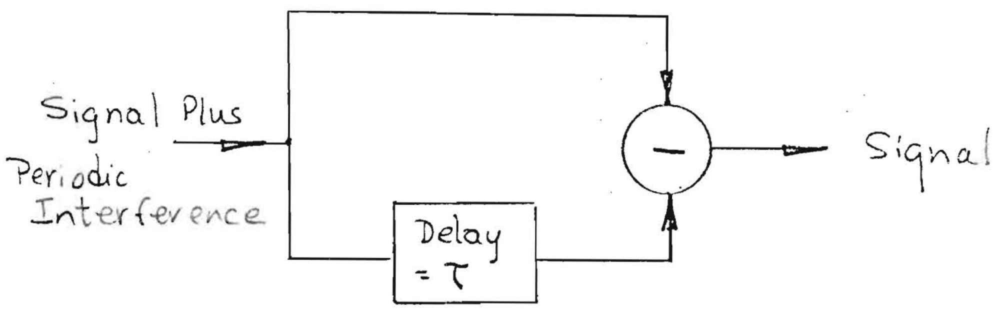



Figure 3 - Simple Filter

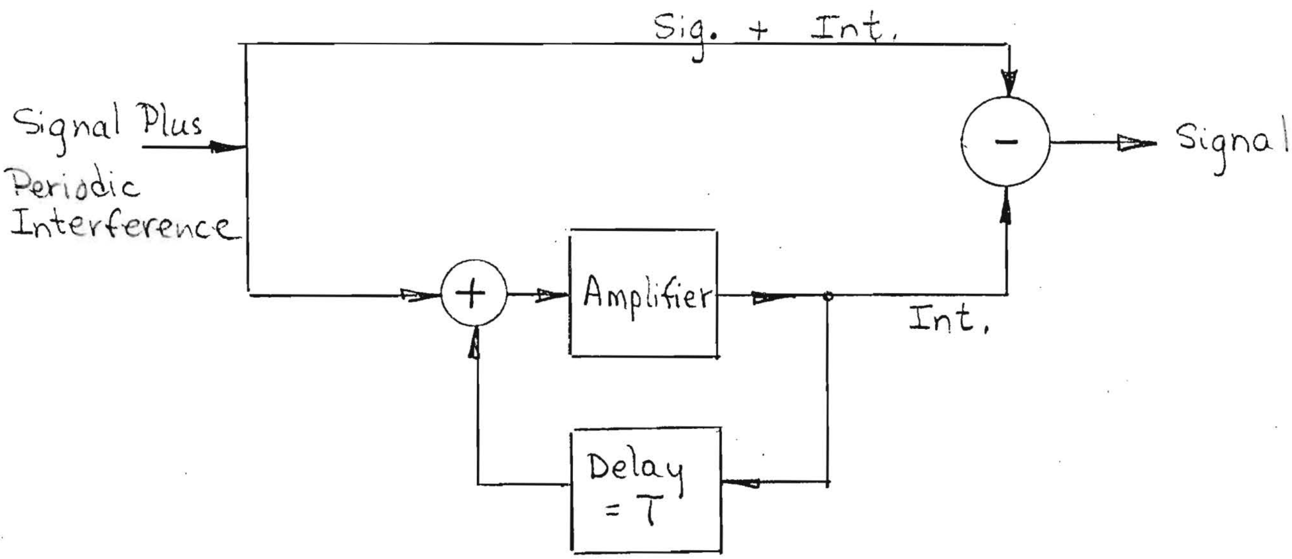



Figure 4 - Improved Filter

GEORGIA INSTITUTE OF TECHNOLOGY
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

17 July 1962

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

Attention: RCUMA

Subject: Monthly Progress Letter No. 19, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

During June the effects of operating point on spurious response rejection for vacuum tube mixers was investigated in some detail. A typical variation of spurious response level with grid bias is shown in Figure 1. From this figure, it can be seen that the spurious response rejection is quite sensitive to this operating point. In view of this sensitivity, it appears that no automatic gain control voltage should be applied to the mixer since this would result in a large variation of spurious response rejection with signal level.

Therefore, no automatic gain control will be applied to either the mixer or the RF stages in the adaptor being constructed. Rather, fixed bias will be used and this bias will be adjusted to minimize the spurious responses of both the RF amplifier and the mixer circuitry.

Physical layout of the cavity preselector, mixer and IF amplifier assembly has been started. Current plans are to provide separate tuning controls for the preselector and the local oscillator circuitry since the design of the mechanical details of the necessary tracking mechanism would be more time consuming than the remaining contract period would permit.

The construction of the delay multivibrator chain for use in the audio filter for rejection of periodic signals has been completed. Tests show that the over-all maximum delay of 1200 microseconds can be obtained. Each delay section is constructed on a printed circuit plug-in card so as to facilitate repair and interchange of delay circuits in case of failure. A total of 24 sections have been constructed.

During the next month, the necessary printed circuit layout and construction of the pulse modulator and demodulator will be undertaken. Since a suitable chassis and power supply for the operation of the entire audio filter have already been procured, the completion of the modulator, the demodulator, and the necessary audio amplifying circuitry will complete the construction of this device.

Finances:

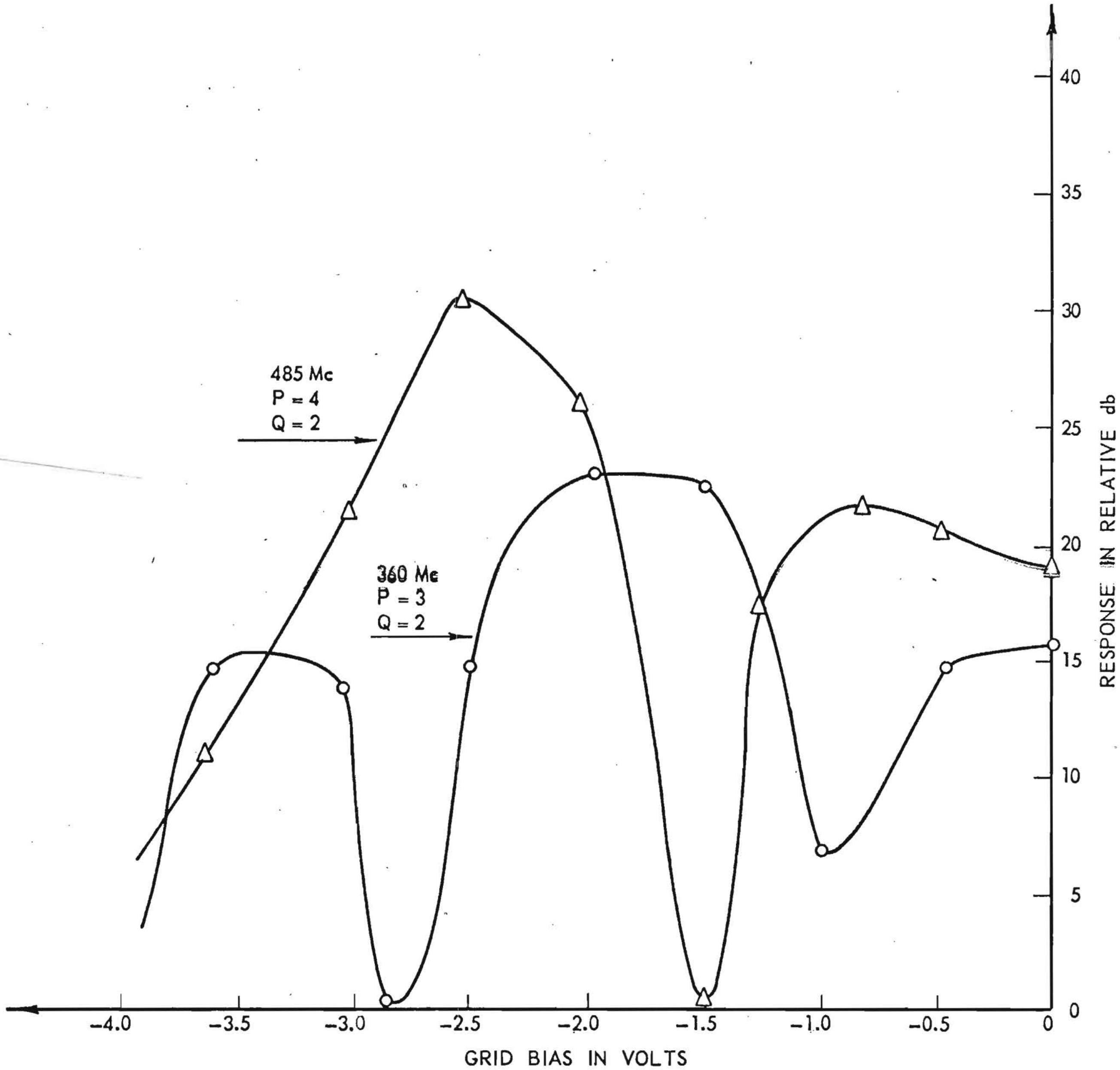
The amount of funds in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

for
W. B. Wrigley, Head
Communications Branch




SPURIOUS RESPONSE LEVEL VERSUS GRID BIAS

Figure 1.

A-525

# GEORGIA INSTITUTE OF TECHNOLOGY

**ENGINEERING EXPERIMENT STATION**

**ATLANTA 13, GEORGIA**

16 August 1962

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

Attention: RCUMA

Subject: Monthly Progress Letter No. 20, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

The RF preamplifier for the receiver front end adapter was completed in July.

This preamplifier consists of two 6J4 grounded grid amplifier stages which produce a gain of about 20db over the 200 to 400 mc band. However, the two tuned circuits necessary for the impedance level adjustment must be tuned separately in order to produce this gain figure. A mechanical tracking arrangement consisting of a gear drive from a single control shaft has been constructed to permit single knob control of both tuned circuits. The tracking error associated with this tuning mechanism reduces the available gain to approximately 17db, but this loss is not considered significant since the net gain is more than adequate to establish a satisfactory noise figure for the entire preamplifier and mixer combination. The added selectivity provided by the two preamplifier tuned circuits is of assistance in reducing the magnitudes of spurious responses at frequencies relatively far removed from the tuned frequency. Addition of this selectivity to that already obtained in the cavity preselector produces approximately 110db rejection on the skirts of the selectivity curve. This rejection, when combined with the inherent spurious response rejection of the mixer, guarantee that the 100db rejection design goal will be achieved, at least for those responses which are relatively far removed from the tuned frequency of the amplifier.

REVIEW

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Some difficulty was encountered in obtaining a 200 to 400 mc tuning range. This arises primarily from the fact that the minimum capacitance values associated with the circuitry limit the available capacity ratio to less than 4 to 1. As a result, a 2 to 1 tuning range could not be obtained by variation of the capacitor alone. To overcome this difficulty, a modified capacitor was constructed with a slider attached to the rotor plates of the variable capacitor. This slider shorts out a portion of the inductance as the tuning shaft is rotated so that a large variation in inductance as well as capacity is obtained. With this arrangement, the desired 2 to 1 tuning range was easily achieved.

Figure 1 illustrates the gain of the amplifier as a function of frequency while Figure 2 illustrates the amount of selectivity obtained in the preamplifier alone. A complete circuit for this preamplifier is shown in Figure 3.

It has been decided that some form of tuning indicator will be necessary in order to make practical use of the adapter. This is primarily due to the fact that no mechanical tracking will be incorporated between the various tuning controls. As a result, some indication of correct tuning adjustment will be necessary in order to simplify the tuning process. To this end two additional IF amplifier stages followed by a detector and an output meter have been added to provide a visual indication of IF signal level. This indication facility will be incorporated as a part of the receiving adapter with the meter located on the front panel.

Progress has been satisfactory in the construction of the delay line audio filter although some difficulty has been encountered with interaction between circuits due to coupling of signals through a common power supply impedance. A few simple changes in the power supply decoupling networks have eliminated this difficulty. Construction of the modulator and demodulator printed circuit boards is almost complete.

Finances:

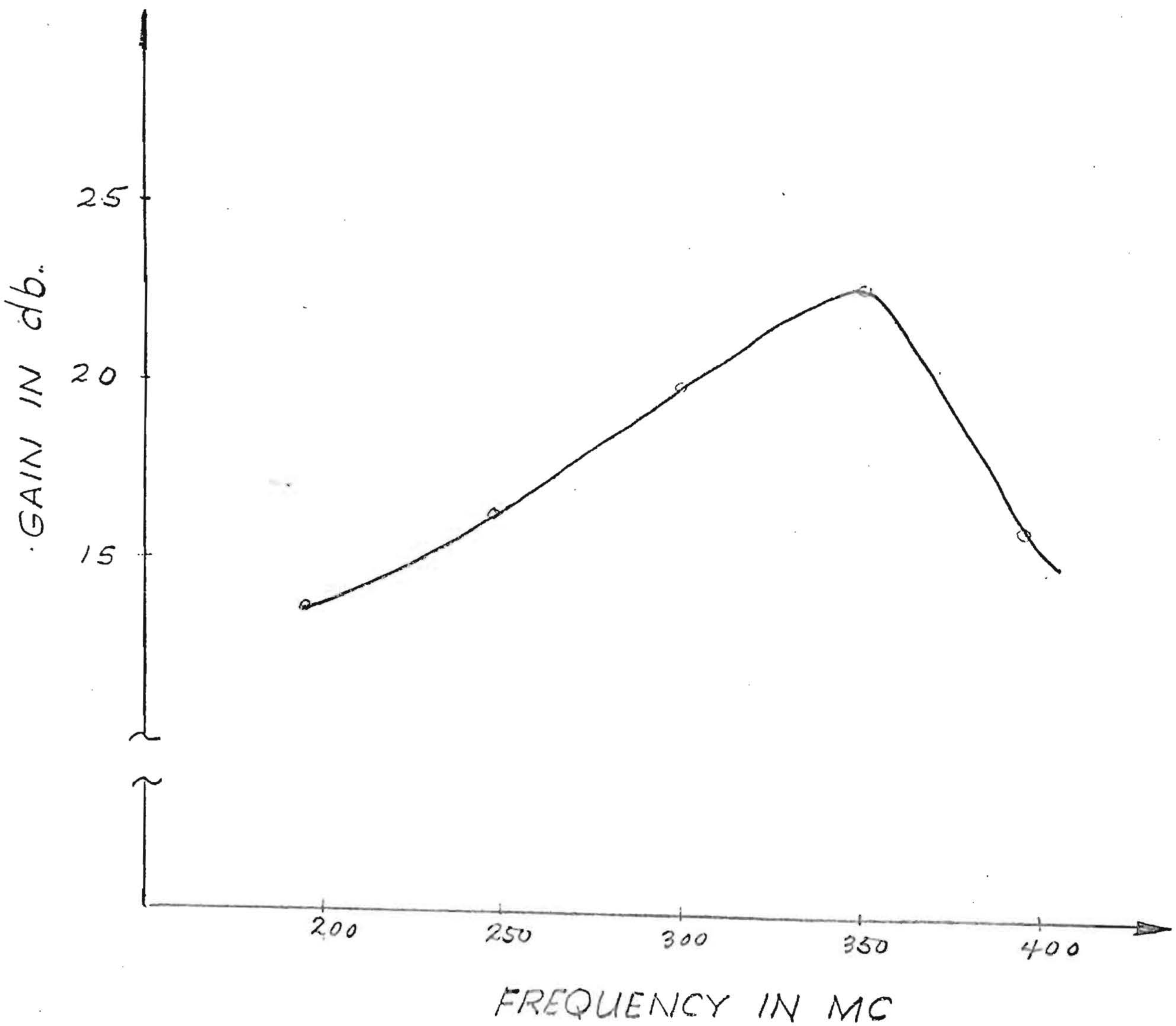
The amount of funds in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

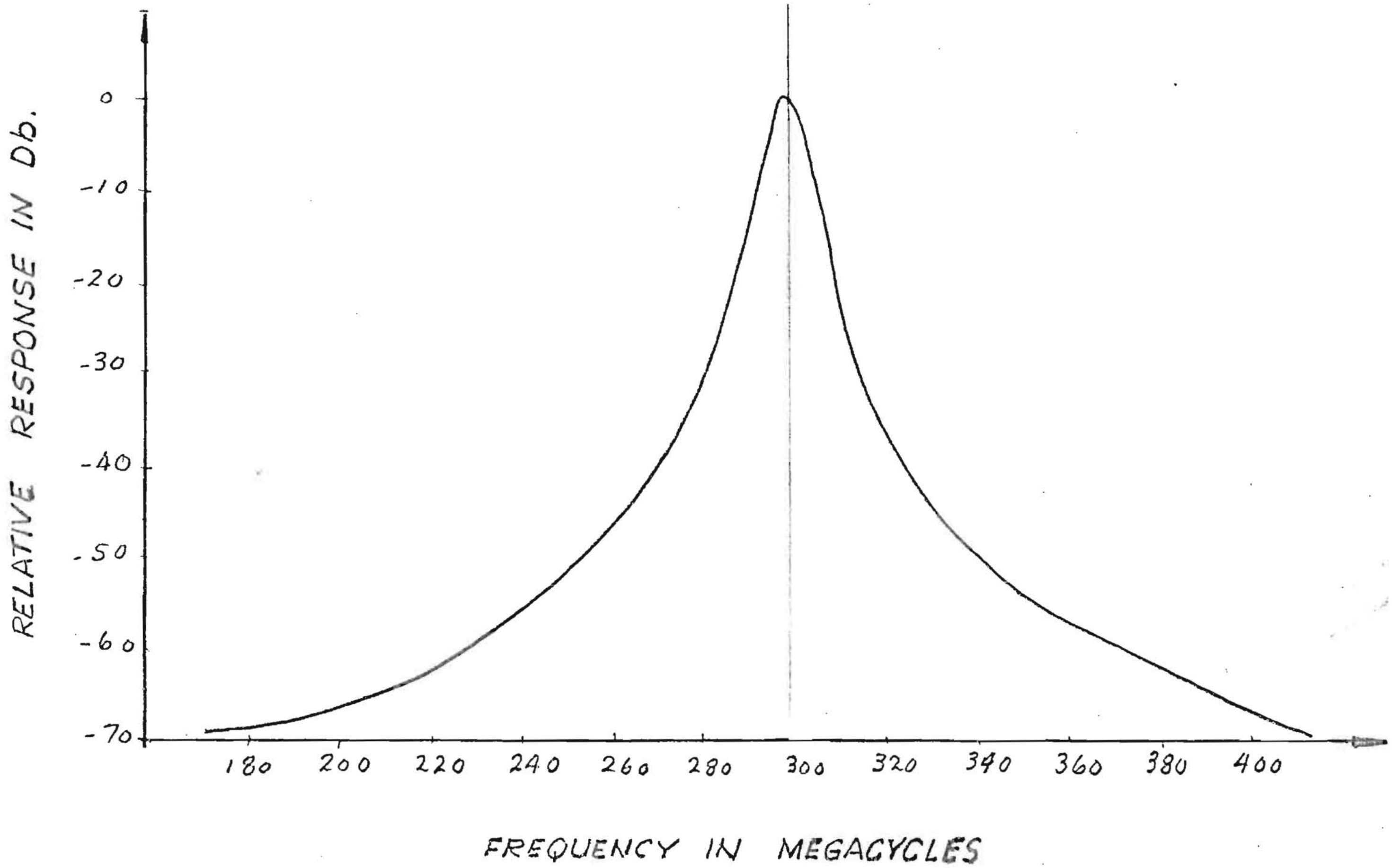
Approved:

for W. B. Wrigley, Head
Communications Branch



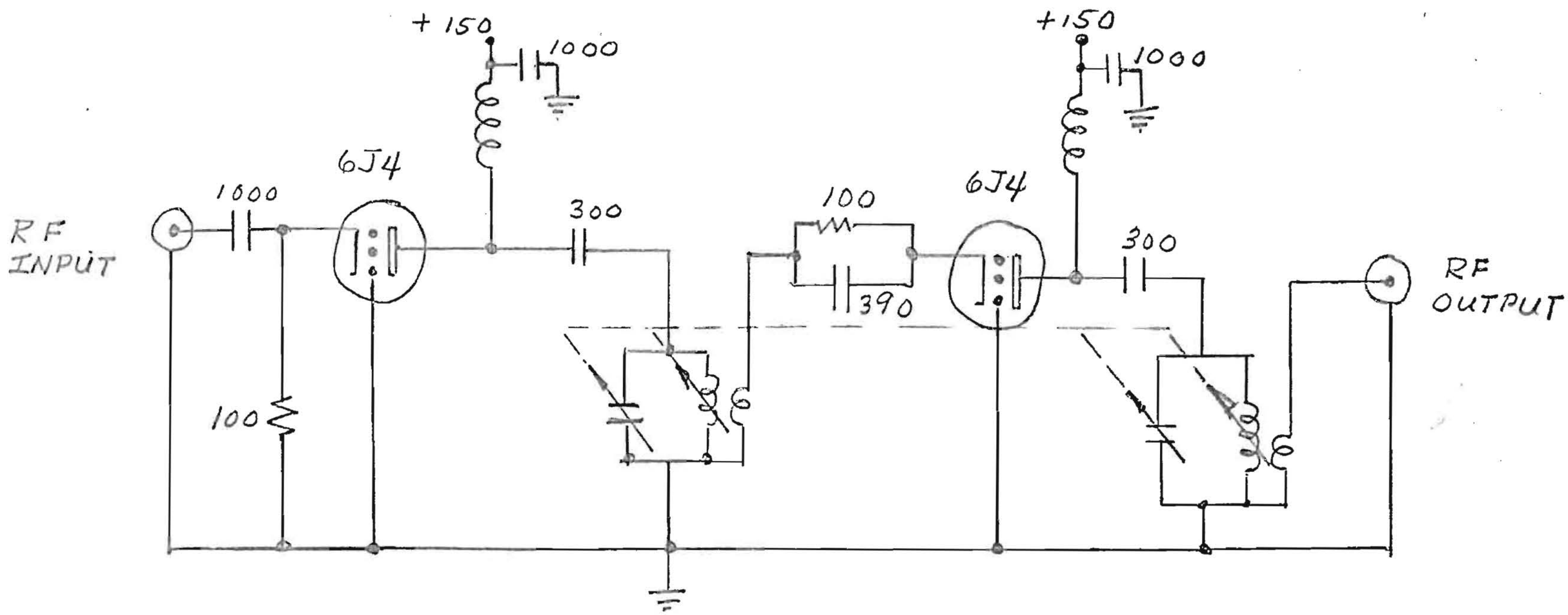

PREAMPLIFIER GAIN VS. FREQUENCY

Figure 1




TYPICAL SELECTIVITY CURVE FOR RF PREAMPLIFIER

Figure 2




RF PREAMPLIFIER

Figure 3

GEORGIA INSTITUTE OF TECHNOLOGY
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

A-525

18 September 1962

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

Attention: RCUMA

Subject: Monthly Progress Letter No. 21, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

During August tunnel diode mixers were investigated for use in applications requiring small, remote preamplifiers. A typical application is that encountered when a measurements antenna is remotely located from a field intensity meter. In such a circumstance, excessive cable losses may attenuate signals below the noise level of the field intensity meter. The location of the mixer directly at the antenna terminals may overcome this difficulty since the cable attenuation may be considerably lower at the IF frequency than at the signal frequency.

Conventional diode mixers have an inherent conversion loss so that the use of such mixers is an advantage only in the circumstance where the conversion loss does not exceed the reduction in attenuation in the cable gained by operating the cable at the IF frequency. For this reason, a tunnel diode mixer offers promise, since its conversion gain may be greater than one.

This conversion gain is a result of time varying conductance of the tunnel diode being negative over a portion of the local oscillator cycle. If this negative conductance is large enough to generate a net negative conductance at the IF frequency, the gain of the mixer can be greater than one.

Figure 1 illustrates the construction of a tunnel diode mixer recently tested. The conversion gain is a function of the local oscillator amplitude since a certain minimum local oscillator signal is necessary to carry the operating point into the negative resistance portion of the tunnel diode characteristic. When the proper amplitude of local oscillator signal was applied, the conversion gain exceeded that of a conventional diode mixer by approximately 15db. The high pass filter at the input prevented the IF signal from being dissipated in the output resistance of the signal source, while the low pass filter at the IF output prevented the incoming signal from being dissipated in the IF impedance. A possible arrangement of a pair of tunnel diodes connected in a balanced arrangement to an antenna is shown in Figure 2. This arrangement permits the local oscillator signals to be fed to both halves of the antenna in phase, so that no radiation of the local oscillator signal from the antenna results. The two out of phase IF components are subtracted in the IF transformer with the result that the IF signal is double that resulting from one mixer alone.

The various components of the receiver adaptor are being assembled in final form for delivery to RADC. Preliminary measurements indicate that the rejection of all spurious responses, with the exception of the image, will exceed 100db. This rejection figure was the target specification originally set for this equipment. The image rejection exceeds 80db. It is anticipated that this equipment will be in complete form for delivery to RADC in late September.

Some difficulties have been encountered in the digital delay chain of the audio filter. This difficulty takes the form of excess pulse jitter when a large number of delay sections are cascaded. Current indications are that this is due to too large a delay in each section so that on peaks of the modulation signal the multivibrator misfires. A reduction in the per-section delay is expected to overcome this difficulty.

Finances:

The amount of funds in the contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

W. B. Wrigley, Head
Communications Branch

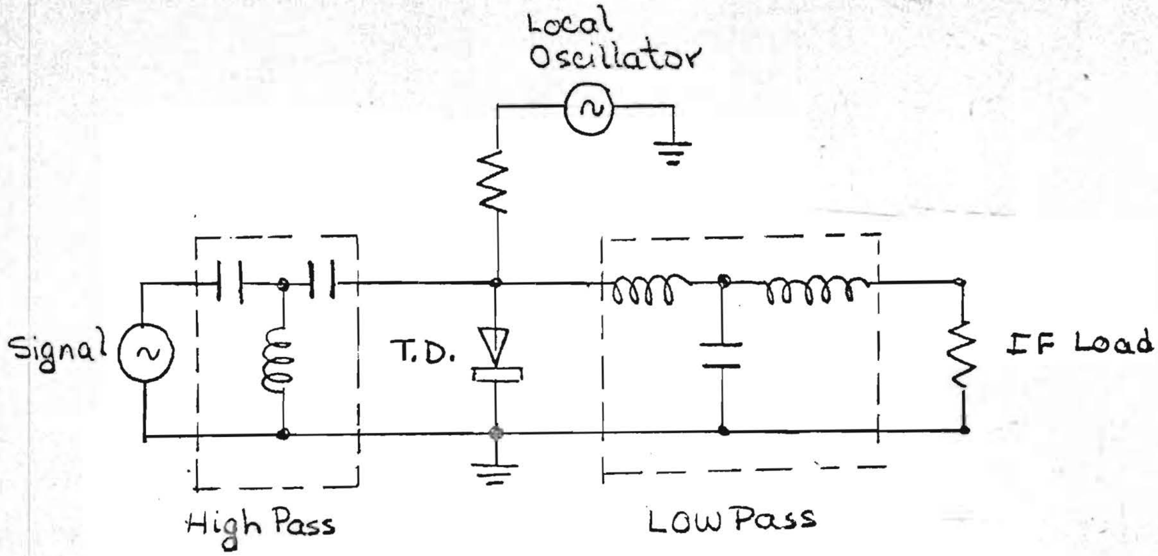



Figure 1 – Tunnel Diode Mixer

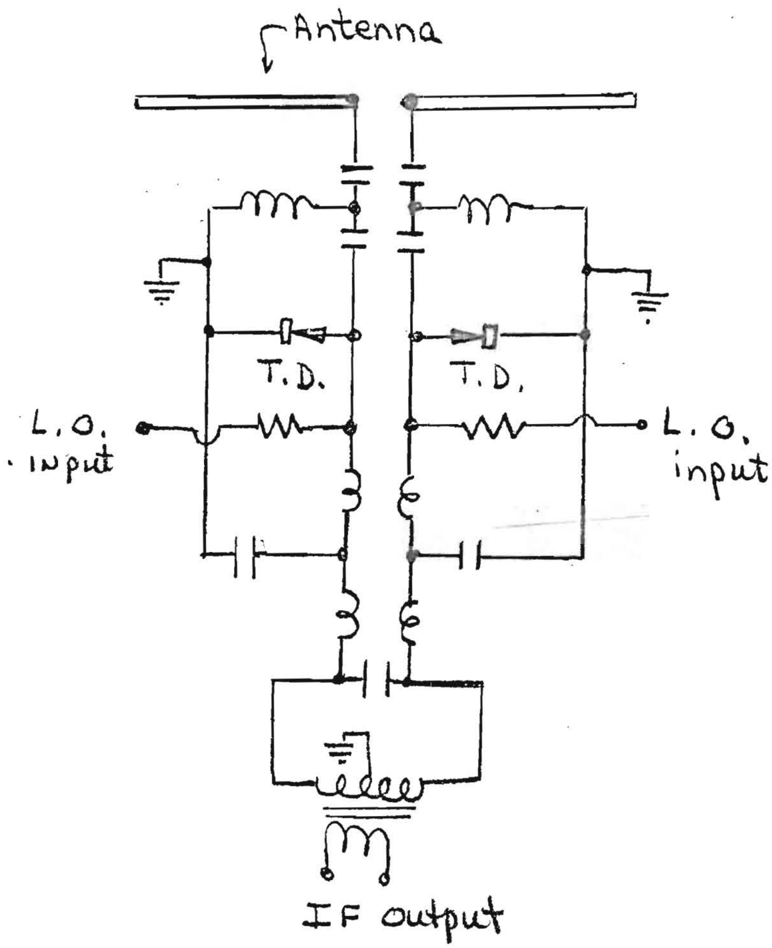



Figure 2 – Balaned Mixer

**GEORGIA INSTITUTE OF TECHNOLOGY**
ENGINEERING EXPERIMENT STATION
ATLANTA 13, GEORGIA

A-575

19 October 1962

Commander
Rome Air Development Center
Griffiss Air Force Base
New York

Attention: RCUMA

Subject: Monthly Progress Letter No. 22, Contract No. AF 30(602)-2366

Dear Sir:

Objective:

To investigate interference suppression techniques to reduce interference from pulsed type emitters.

Technical Program:

The difficulty encountered in the digital delay chain of the audio filter mentioned in the Progress Letter for August has been overcome. The excessive jitter, which occurred when a large number of delay sections were cascaded, resulted from the fact that the delay of the multivibrator depended upon the time between trigger pulses and, hence, would vary with the signal modulation. This sensitivity to signal modulation was the result of incomplete recharging of the timing capacitor used in these delay multivibrators. Each multivibrator has been modified to include an additional emitter follower which provides a very low impedance recharging path for the timing capacitor. The method by which this emitter follower has been incorporated in the basic delay multivibrator is illustrated in Figure 1. The inclusion of this modification has resulted in a delay chain which is free of jitter. When this delay line was connected in the configuration shown in Figure 2, good cancellation of a periodic signal was attained.

During September the receiver adaptor was assembled in final form and delivered to RADC. This equipment includes a combination of several techniques for reducing spurious responses in communication receivers and equipment. The specific items included are: 1. a cavity preselector; 2. a linear RF amplifier; 3. a low spurious response balanced mixer; 4. a local oscillator system in which the harmonic output has been drastically reduced.

These separate devices have been arranged in such a manner that external connection may be made to each of them by means of front panel jacks.

As a result, any desired combination of these devices can be used to illustrate the improvement in spurious response rejection obtained in a given situation.

Work has begun on the draft of a final report covering the work under the contract. No difficulties are foreseen in completing this effort in the time allotted between the end of the technical effort on October 15, and the November 15, date required for submission of the report.

Finances:

The amount of funds in the current contract is sufficient to cover the present and anticipated rate of expenditure.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

D. W. Robertson, Head
Communications Branch

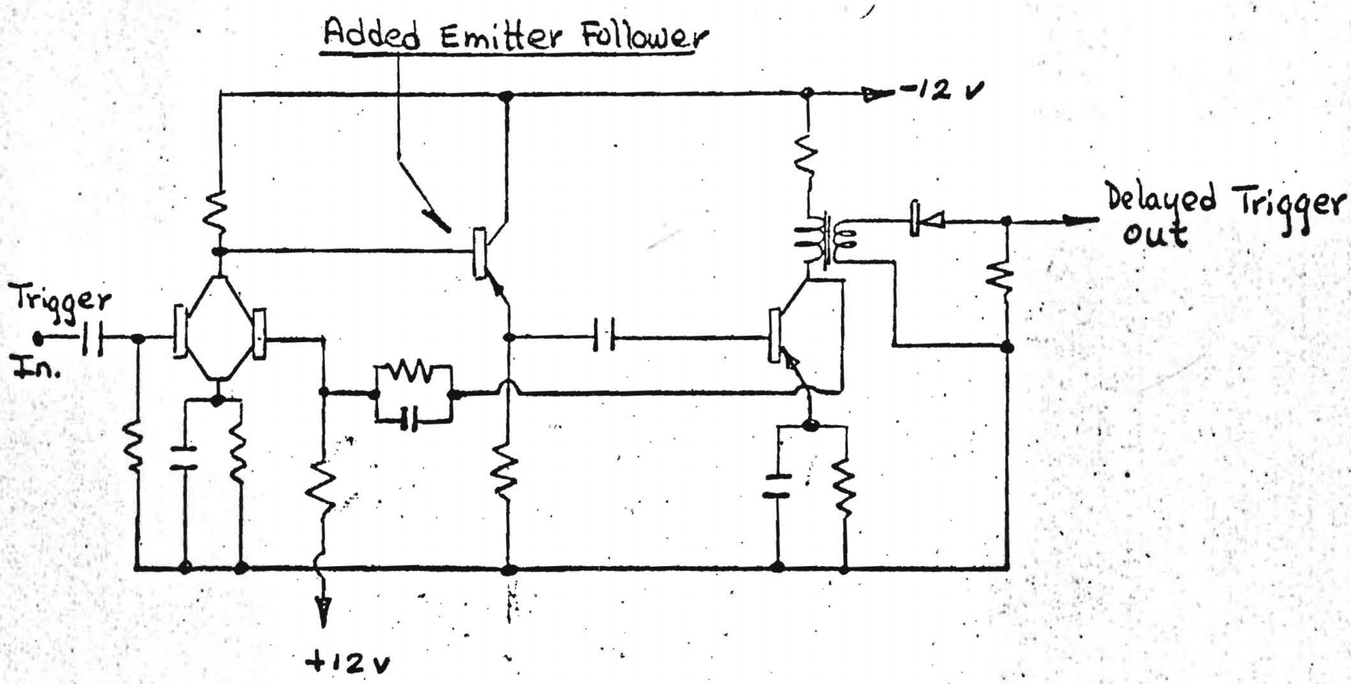



Figure 1- Modified One-Shot Multivibrator

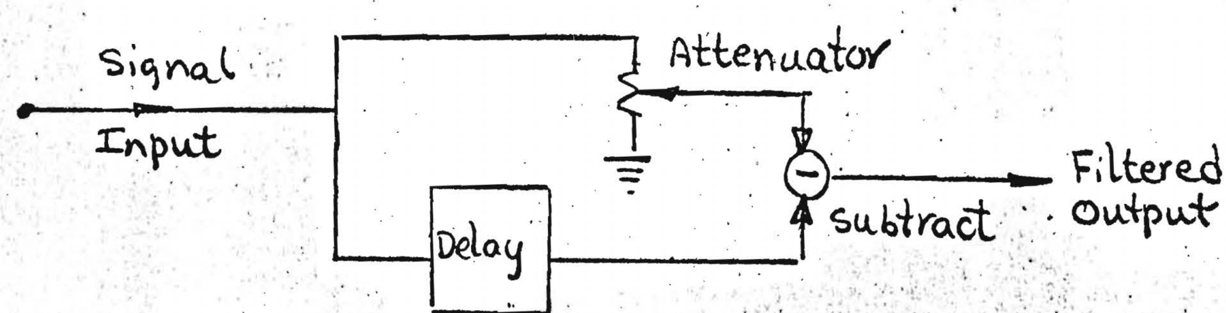



Figure 2 - Audio Filter

RADC-TN-61-330

# A PULSE INTERFERENCE BLANKER

By

W. B. Warren, Jr.

TECHNICAL NOTE NO. 1

Contract No. AF 30(602)-2366

Project No. A-525

Prepared for

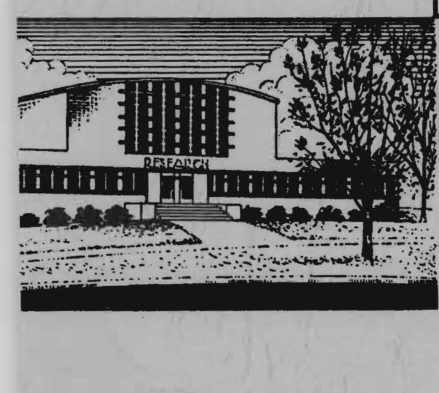
Rome Air Development Center

Air Research and Development Command

United States Air Force

Griffiss Air Force Base, New York

15 November 1961



Engineering Experiment Station

Georgia Institute of Technology

Atlanta, Georgia

RADC-TN-61-330

# A PULSE INTERFERENCE BLANKER

By

W. B. Warren, Jr.

ENGINEERING EXPERIMENT STATION
of the Georgia Institute of Technology
Atlanta, Georgia

TECHNICAL NOTE NO. 1

Contract No. AF 30(602)-2366

Project No. A-525

Prepared for
Rome Air Development Center
Air Research and Development Command
United States Air Force
Griffiss Air Force Base, New York

15 November 1961

## ABSTRACT

An interference blanker is described which is effective in overcoming the effects of pulse interference to narrow band communication receivers. This device employs both the techniques of "blanking" and "sampling" to obtain the desired interference suppression.

Explanations and mathematical justification for both modes of suppression are presented. A description is given of a completed device which permits the protection of communications equipment without the necessity of making internal modifications.

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# 1. INTRODUCTION

In many instances, it is necessary to co-locate narrow band communications equipment with high-powered, pulse-type equipment, such as radar. In this situation, it is not always possible to prevent a considerable amount of pulse energy from appearing at the input to the communications equipment. In some instances, this power level may be as high as several watts. Such power levels, incident upon conventional communications equipment, will cause serious overloading and desensitization and render impossible the reception of weak desired communications signals. This difficulty is due to the fact that most communication receivers do not provide sufficient frequency selectivity in their input stages, thereby permitting considerable amounts of the high-powered pulse interference to reach the grid circuits of the first few stages in the receiver. Overloading of these stages by the pulses causes grid current to be drawn with a consequent rapid charging of the AVC line and severe desensitization of the receiver because of the excessive AVC voltage which is developed. Consequently, pulse interference suppression techniques which are applied at a point in the receiver which follows these input stages are not effective in suppressing the effects of this interference, since the damage has already been done before the signal reaches the point where the suppression device is applied. Adequate suppression of the effects of this type of interference can be obtained only if the interfering pulse is rejected directly at the input terminals of the receiver.

It is the objective of this Note to describe a device which permits the suppression of the pulse interference directly at the input terminals to a communication receiver so as to avoid the difficulties mentioned previously. In addition, this device requires no internal modification to the receiver with which it is being used.

## 2. CONCLUSIONS

It is concluded that excellent protection of communications equipment from pulse interference can be obtained by the use of devices similar to the interference blanker described in this Technical Note. This device has been packaged in a form suitable for use in field testing, and its operation has been successfully demonstrated at RADC.

# 3. RECOMMENDATIONS

The results obtained in the application of the interference blanker to typical pulse interference situations indicate that this type of equipment may be successfully applied whenever problems of pulse interference to communications equipment are encountered.

It is recommended that this device be field tested in a variety of pulse interference situations to determine its applicability to a wide range of actual interference conditions.

## 4. DISCUSSION

In this device, called an "Interference Blanker", as is true with any other interference suppression device, it is necessary to recognize some essential difference between the desired and interfering signals and to make use of this difference to obtain the desired interference reduction. In the case of a narrow band desired signal, which is modulated with speech or other narrow band intelligence and which is being interfered with by a pulse-type signal, one essential difference lies in the fact that the interference is limited in time, while the desired signal is not. Thus, by turning the input to the narrow band receiver on or off at the proper instants of time, the interfering signal can be drastically reduced in amplitude or completely eliminated. On the other hand, the desired signal, not being of a time limited character, is sufficiently unaffected by this gating of the receiver input to permit the desired signal to be successfully recovered.

4.1. Blanking. The Interference Blanker applies the gating technique in either of two basic ways. The first of these is called "blanking". In the blanking technique, the receiver is turned off for the duration of each interfering pulse and turned on again between pulses, but the gaps in the desired signal due to this gating process do not seriously degrade the intelligibility of the desired signal. The blanking is accomplished by the insertion of a pulse controlled switch in series with the input to the communication receiver being protected, in the manner shown in Figure 1. The action of the pulse controlled switch in removing the interfering pulse signal is indicated by the wave forms of Figure 2, which illustrate the situation of a desired, sine wave modulated AM signal with a periodic pulse interference superimposed. Notice that this technique will be effective only if the blanking pulses are synchronous with the interfering pulse

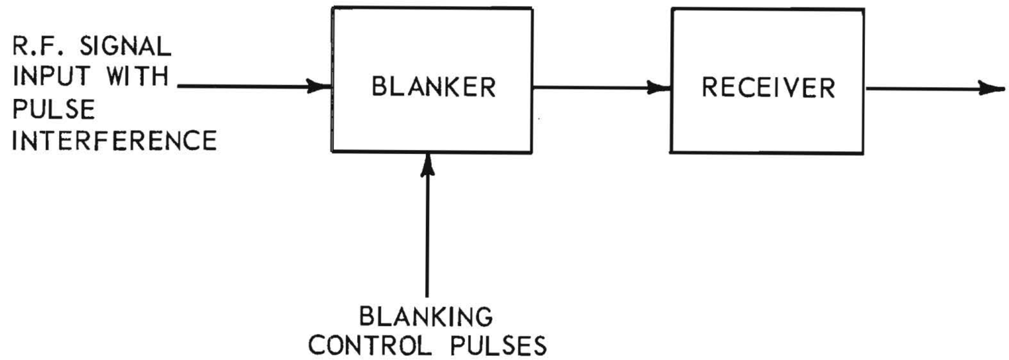



Figure 1. Interference Blanking

signal. In cases where it is possible, the best method to obtain this synchronization is to have a direct cable connection with the source of interference so that a pre-trigger for the blanking pulse can be obtained. The pre-trigger is necessary, since there is some delay in the generation of the blanking pulse and in the disconnect action of the blanking switch. As a result of this delay, it is not possible to blank a particular interference pulse by detecting that particular pulse and using this detected output to generate the necessary control pulse for the blanking switch unless an adequate delay in the main signal path can be obtained. In general, such a delay in the signal path is difficult to obtain and has the additional disadvantage that most devices which might be used for this purpose have an appreciable insertion loss so that the desired signal would be unduly attenuated in passing through them. However, if the proper pre-trigger pulse is available, the blanking switch may be operated by this pulse slightly before the arrival of the interference pulse at the receiver input and complete blanking of the interfering signal is obtained. If a direct cable connection to

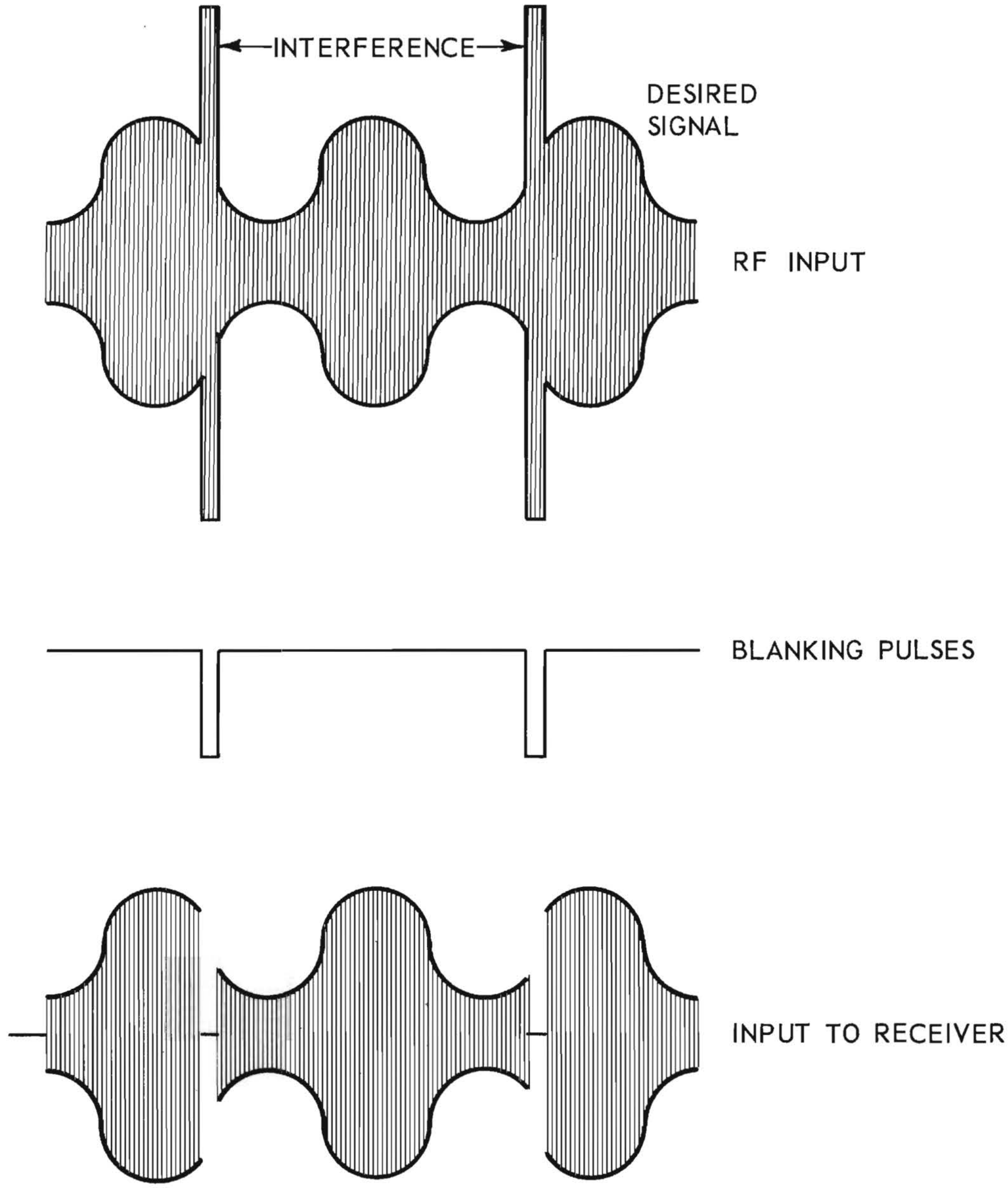



Figure 2. Blanking Waveforms.

the source of pulse interference is not available, an auxiliary receiver can be used to detect the interfering pulses and supply them to a blanking pulse generator as synchronizing information.

In the latter situation, it is not possible to have the proper pre-triggering information, due to the previously mentioned delay, which is associated with the switch and will also be associated with the auxiliary receiver. However, for periodic interference, the required pre-triggering information can be obtained by delaying the interfering pulses by an amount just short of one pulse interval. In this arrangement, the blanking pulse corresponding to a particular interfering pulse is generated by the previous interfering pulse delayed by almost one pulse interval. This technique is effective as long as the condition of periodicity is met.

If the multiple interference sources are encountered and if the rates of these sources are not integrally related, it is not possible to gate out all the interference. In fact, if the blanking pulses are synchronized to one of the interfering sources, then every pulse of a non-related interfering source will enter the receiver, except those which overlap the pulse interval of the source to which the blanking pulses are synchronized.

4.2. Sampling. A second receiver gating technique which is used in the Interference Blanker to reduce or eliminate pulse interference is called "sampling". In this process, samples of the input signal are taken at a rate which is sufficiently high to reconstruct all the essential information in the desired signal from knowledge of the samples alone. Since the required information in the input signal is the envelope (in the case of an AM signal), it is not necessary to provide samples at a rate high in respect to the carrier frequency of the input signal, but rather it is sufficient to provide samples at a rate which is high with respect to the highest frequency components contained in the envelope of the desired signal. For the case of a speech modulated signal occupying the frequency range 200 to 4,000 cps, the required minimum sampling rate is 8,000 per second, in accordance with well known theory of sampled signals. However, in a practical situation, the perfect filters required by the theory are not available and a slightly higher sampling rate is necessary to permit "clean" recovery of the original signal from knowledge of the samples alone. For the case mentioned above, a practical sampling rate of the order of 10,000 samples per second is necessary. Since the sampling of the input signal is done at the RF input to the receiver, it is necessary to reconstruct the desired signal from the sampled signal by means of a band-pass filter, rather than by a low-pass filter, as is usually the case.

Figure 3 illustrates the technique of sampling an amplitude modulated signal and reconstructing the original signal by a band-pass filter. Since the output of the sampling switch is zero, except at the sampling instants, it is apparent that the output of the band-pass filter is dependent only on the values of the input AM signal at these sampling times, since the input to this filter is zero at all other times. Hence, any interfering pulse signals which might occur at times other than the sampling instants will not be present in the output of the sampling switch and, therefore, will not be present in the output of the band-pass filter.

In a practical situation, the bandwidth of the RF amplifier portion of a conventional receiver is not sufficiently narrow to act as a perfect band-pass filter, such as is assumed to be available in acquiring the wave forms of Figure 3. As a result, the input to the mixer and, hence, the input to the IF amplifier

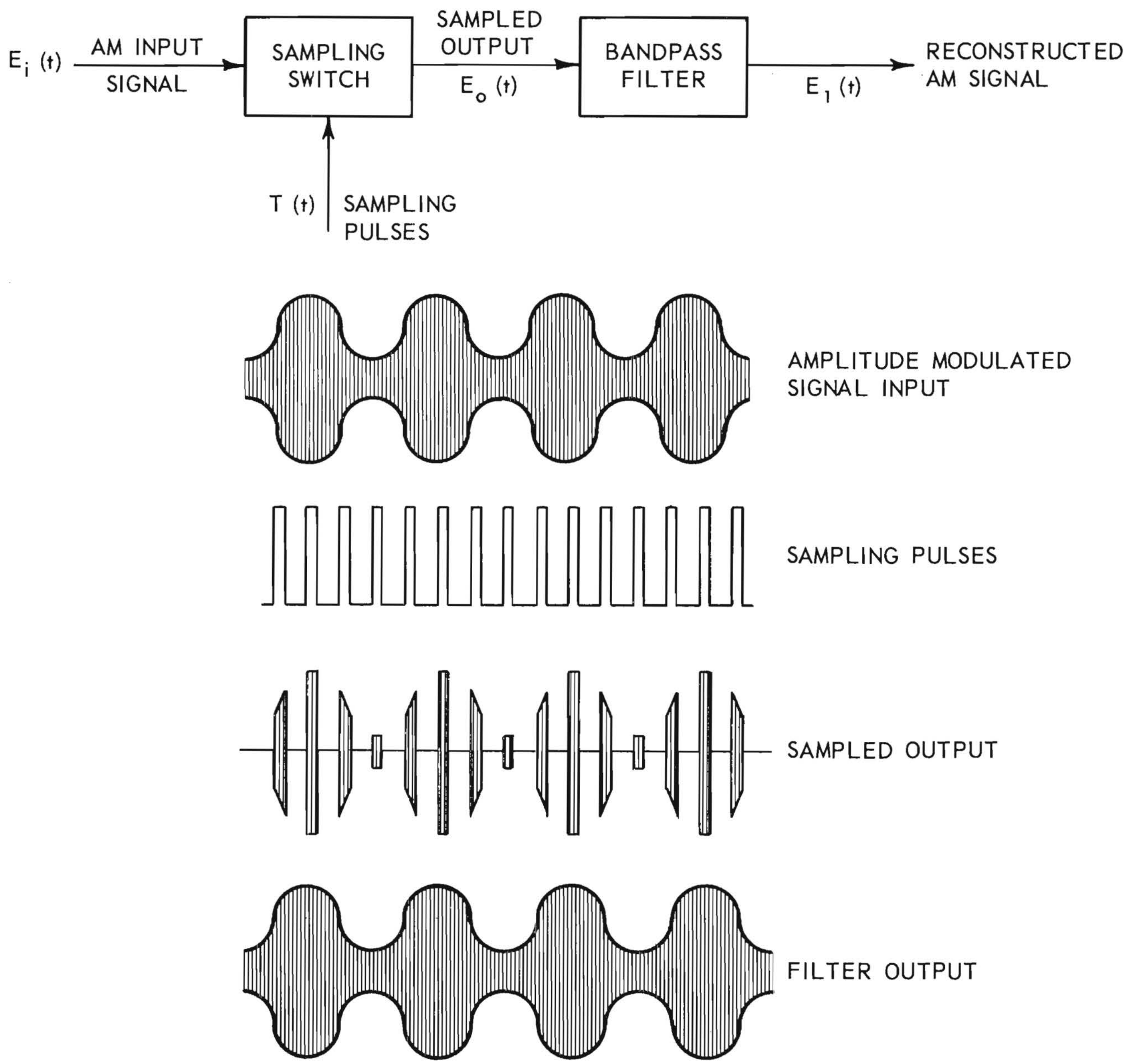



Figure 3. Sampling Technique.

is in the nature of a pulse amplitude modulated signal, similar to that labeled "sampled output" in Figure 3, with the exception that the carrier frequency of this pulse amplitude modulated signal is now the IF frequency of the receiver, rather than that of the original input signal. The narrow bandwidth of the IF amplifier then serves as the required band-pass filter so that its output will appear as the reconstructed AM signal and as such may be applied to an envelope detector to produce the desired envelope modulation at audio frequency.

4.2.1. Spurious Responses. A mathematical justification of the fact that this sampling technique does indeed preserve the envelope modulation on the desired signal is given in the Appendix. In this analysis, it is pointed out that the process of sampling constitutes a modulation of the input signal at the sampling rate, and hence produces sidebands which are distributed in frequency on both sides of the desired signal frequency. This is an undesirable situation, since it causes the receiver to have a spurious response at frequencies close to the frequency of the desired signal. The magnitude of these responses is governed by the magnitude of these sidebands, which in turn are functions of the shape of the sampling pulses. As a result, it is possible to reduce the level of these spurious responses considerably by careful attention to the shape of the sampling pulses. Nevertheless, this production of spurious responses represents a disadvantage in the use of the sampling technique. In addition, it should be noted that the desired signal source is connected to the receiver input for a small fraction of the total time; this fraction being equal to the duty cycle of the sample pulses. However, the receiver noise as generated in the input stages of the receiver is present all the time, with the result that the signal to noise ratio of the input signal is a function of the duty cycle of the sampling process. It is apparent, then, that for a constant sampling rate, the width of the sampling pulses should be made as wide as possible. On the other hand, the probability that interfering pulses will occur at one of the sampling instants is directly related to the width of the sampling pulses. Therefore, a compromise must be reached in which the maximum pulse width is used which is consistent with the desired degree of interference reduction.

4.2.2. Calculation of Interference Reduction. Some measure of the degree of interference reduction that may be obtained by the use of a sampling scheme such as is described here may be obtained by considering the probability that an interference pulse occurs during one of the sampling times. If there is synchronization between the sampling pulses and the interfering signal, then there is no matter of probability involved, since the existence or absence of interference depends on the phase relationship between the sampling pulses and the interfering signals and may be calculated explicitly. However, for those cases where the sampling pulses occur in a somewhat random fashion with respect to the interfering pulses, the question of the absence or existence of interference reduces to the determination of the probability that an interference occurs during the time of the sampling pulses.

At any instant of time, the probability that a sample pulse occurs is given by the duty cycle of the sample pulse train. Likewise, the probability of occurrence of an interfering pulse is given by the duty cycle of the interfering pulse train. If these two events are statistically independent, then the probability of simultaneous occurrence of these two events is given by the product of their probabilities. Thus, if

$$f_1 = \text{sampling pulse rate}$$

$f_2$ = interfering pulse rate

$\Delta_1$ = sampling pulse width

$\Delta_2$ = interfering pulse width

then the per cent of time that coincidence occurs is given by

$$\% \text{ coincidence} = (\Delta_1)\ (\Delta_2)\ (f_1)\ (f_2). \qquad (1)$$

This particular mode of operation, then, is best suited to those situations where multiple, asynchronous, interference sources are encountered. Since the sampling pulses may be synchronized so that samples are taken immediately before or after the occurrence of the pulses of one of the sources, the number of the interfering pulses from other sources, which will enter the receiver, will then be reduced by the factor given in Equation (1).

4.3. Diode Switch. In the application of either the blanking or sampling techniques to the reduction of pulse interference, it is necessary to have a switch which can be controlled by the blanking pulses to effect a rapid disconnection of the receiver from its antenna so as to prevent interfering pulses from entering the receiver. At the same time, this switch must be capable of a very low minimum insertion loss when it is used to connect the receiver and the antenna together so as not to unduly attenuate low-level desired signals. The basic design of such a switch, capable of operation in the frequency range of 200 to 400 mc, has been developed by Georgia Tech under Contract No. AF 30(602)-1789. This switch is essentially a multi-section low-pass filter whose cut-off frequency is about 400 mc. Each section of this low-pass filter has a normally back-biased diode connected in shunt with it so that the effect of the diode is to add a small amount of shunt capacity to each section of the filter. This small added capacitance is taken into account in the design of the filter so that the effect of the back-biased diode is essentially negated and the attenuation in the pass band of the filter is quite small. However, the application of a forward biasing voltage to the shunt diodes causes them to exhibit a very low shunt resistance across each section of the filter, and the attenuation of the filter becomes very large. This large attenuation results from two mechanisms. The first is the considerable impedance mismatch encountered at the input to the filter when the diodes are forward biased, with the result that only a small fraction of the incident power actually flows into the switch proper; the larger portion of the power being reflected back to the source. That fraction of the power which does enter the filter is then attenuated in each successive section of the filter by an amount dependent upon the ratio of the shunt diode resistance to the series reactance of the inductive arm of that particular section.

This mechanism may be visualized by referring to Figure 4 which shows, schematically, the construction of the switch. Two capacitors are used at the input and output of the filter to isolate the filter for DC; however, the reactance of these capacitors at frequencies in the 200 to 400 mc range is negligible. The control pulse is applied to the filter through an RF choke whose impedance is high with respect to the 50 ohm impedance level of the filter but does not present any appreciable reactance at frequencies contained in the control pulse. The details of the mechanical construction of this filter can be seen in the photograph of Figure 5. The series inductance is provided by short lengths of No. 10 wire, while the shunt capacitance is supplied by silver mica button capacitors. This particular type of capacitor was used because of the low inductance connection which can be made to it, and because it concentrates its capacitance at one point. These button capacitors also serve as feed through connections

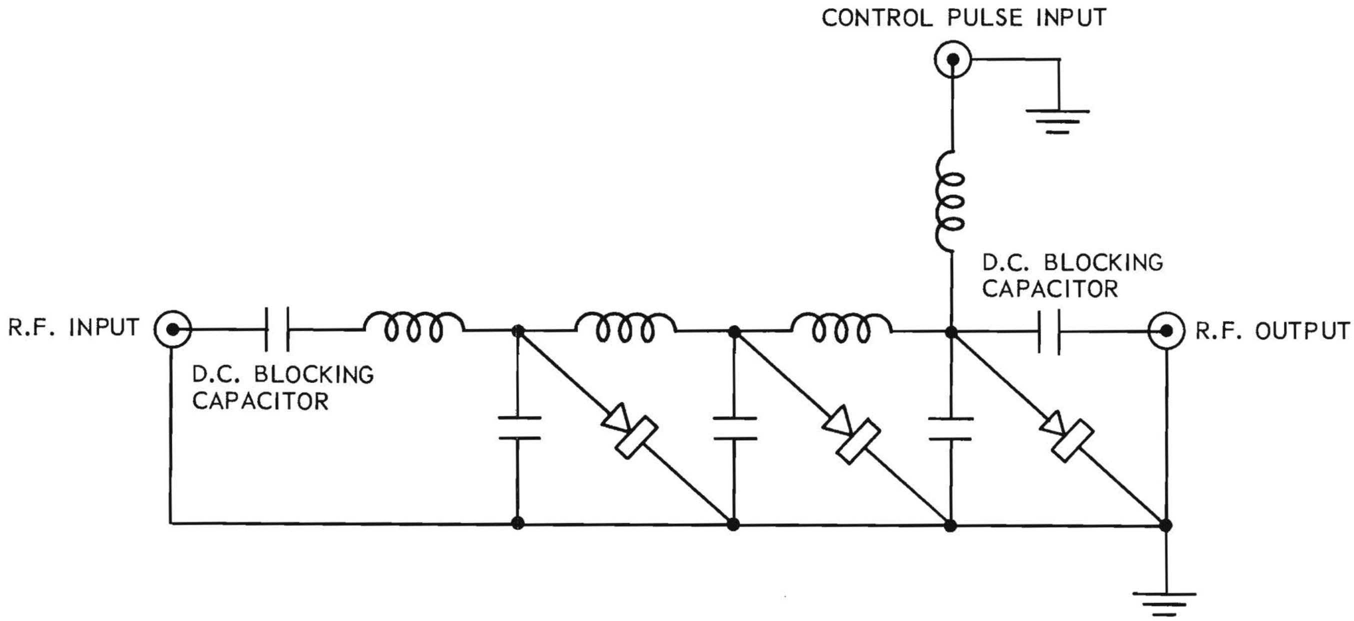



Figure 4. RF Switch Schematic

between adjacent sections of the filter. The diodes are connected between the center conductor and ground and are physically located at the point where the center conductor connects to the shunt capacitors. The entire assembly is constructed in a box approximately one inch in cross section and eight inches long. The 50-ohm impedance level of the filter is sufficiently low with respect to the impedance of the box viewed as a short length of transmission line so that the box introduces no appreciable effect on the response of the filter.

A set of typical frequency characteristics of a diode controlled switch of the type shown in Figure 5 is shown in Figure 6. The insertion loss with a 400 milliampere control current exceeds 56 db over the range of 200 to 400 mc, while the insertion loss with the diodes back-biased is less than 1 db over the same frequency range. This 56 db insertion loss is sufficient to reduce a 10 watt

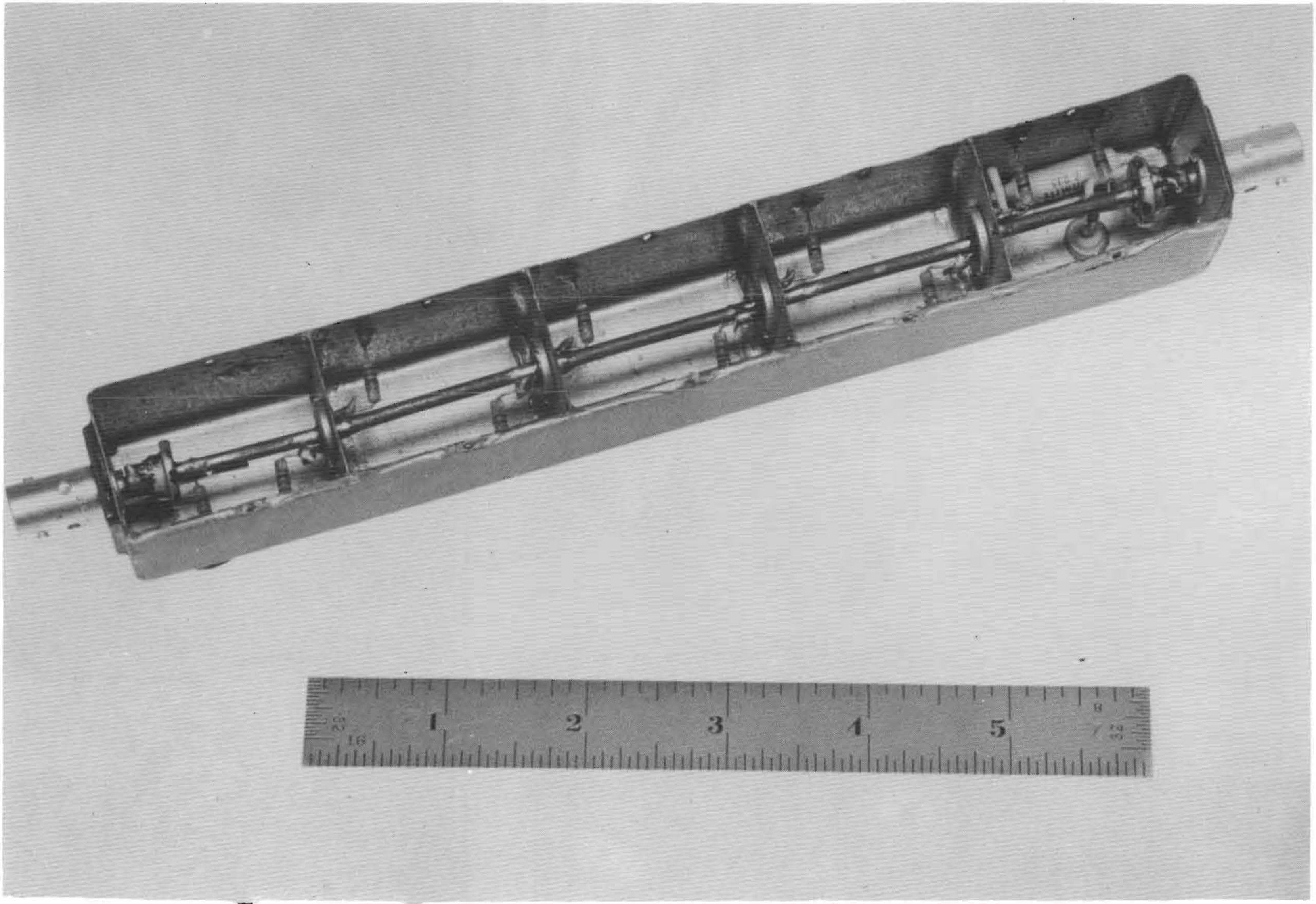



Figure 5. Photograph of Diode Switch.

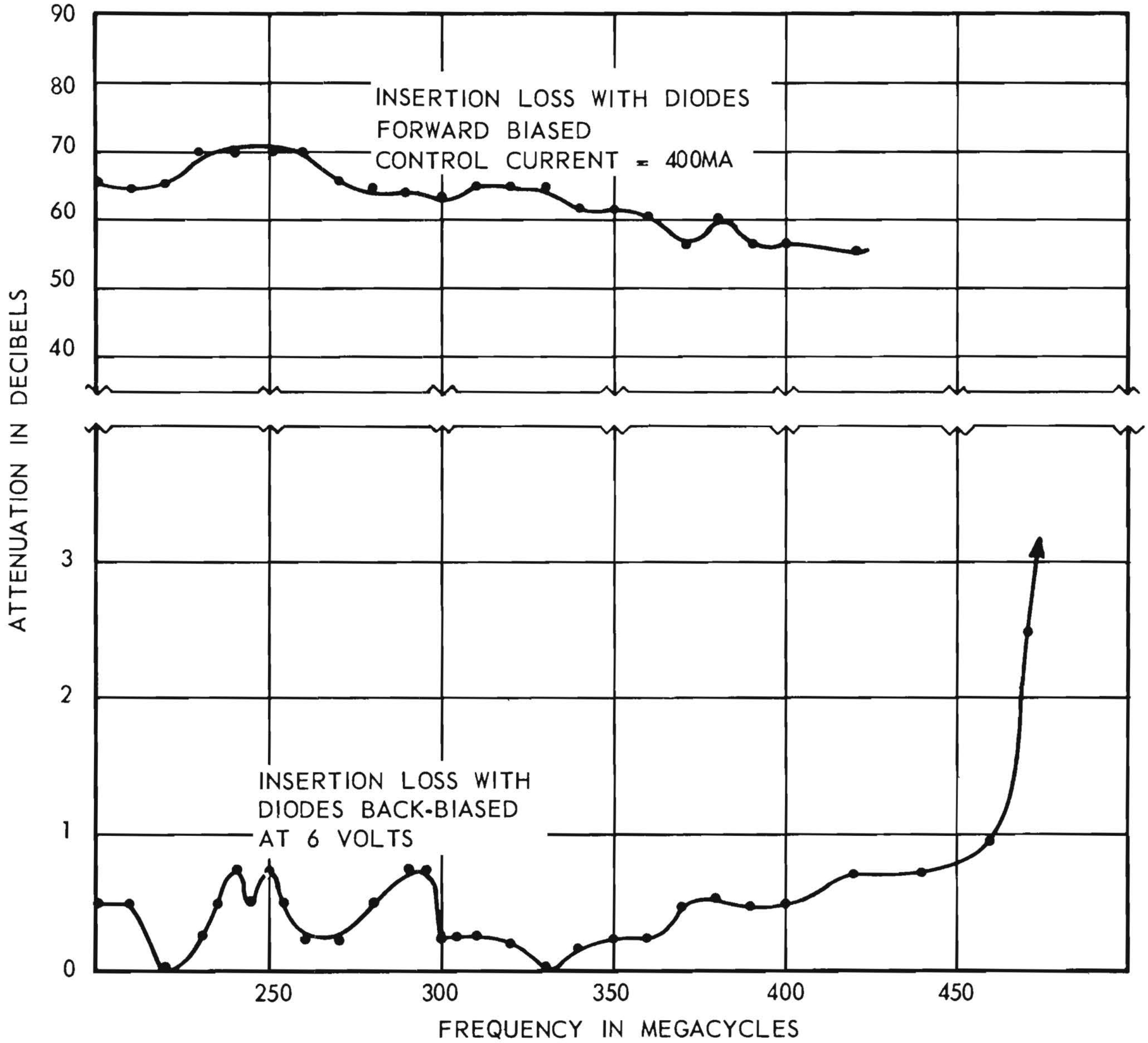



Figure 6. Switch Insertion Loss Characteristics.

interfering pulse appearing at the antenna terminals of the receiver to a level of $25 \times 10^{-6}$ watts. Although this level is quite low and will not cause serious overloading in the front end of the receiver, it is still sufficiently large to cause considerable difficulty in the reception of desired signals close to the sensitivity threshold of the receiver. It is also possible that this remaining level is large enough to produce a considerable AVC voltage at the output of the AVC detector, especially in those receivers having conventional peak detecting AVC rectifiers. This excessive AVC voltage may cause sufficient desensitization of the receiver to prevent the reception of the very weak desired signals. In such an event, two of the diode switches may be connected in cascade so that an additional 56 db of attenuation of interfering pulse signals is obtained. This amount of attenuation is sufficiently large to reduce the level of the interfering pulses almost to the noise level of the receiver. Even with this arrangement, the maximum insertion loss in the back-biased condition is considerably less than 2 db over the entire range of interest from 200 to 400 mc. With regard to the cascade connection of two of these switches, care must be taken to connect the control pulse inputs of the two switches in a "back to back" manner. This is illustrated in Figure 7. This is necessary because the leakage of the signal out of one of the switches into the common control pulse lead can effectively by-pass the action of the second switch when the "back to back" connection is not used.

In the use of a switch of this nature, there is a change in the current in the inductance of the low-pass filter, when the diodes are switched from the forward to back-bias condition. Since a portion of this current flows in the output circuit of the switch, it is possible that extraneous signals may be presented to the input to the receiver even though no input signal is incident upon the

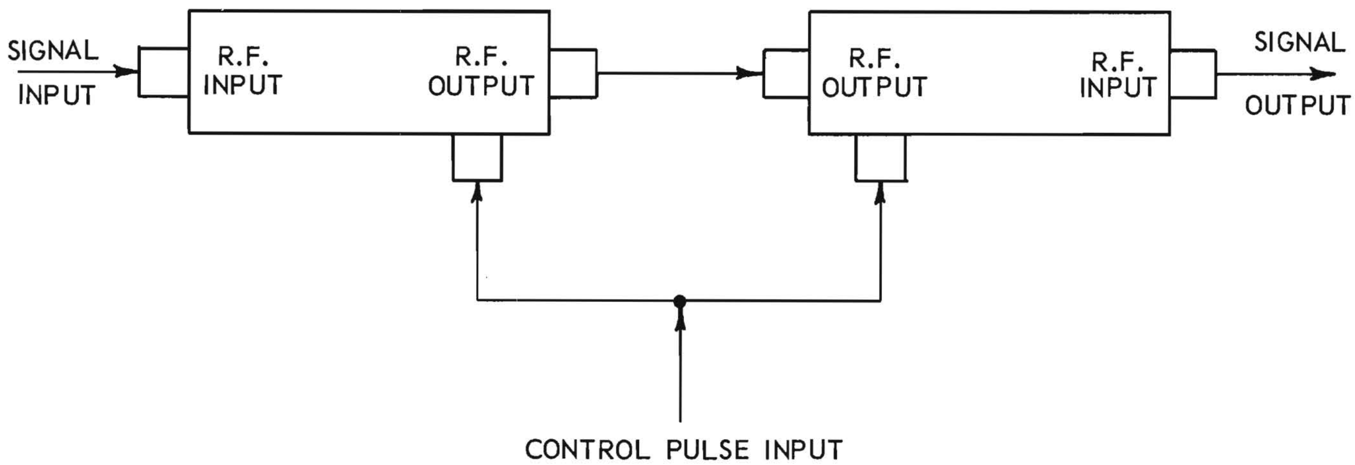



Figure 7. Proper Method of Cascading Two Diode Switches

antenna. Fortunately, this difficulty is not encountered in the use of this switch in the frequency range 200 to 400 mc, since the switching of the diodes from the forward to back-bias condition is slow enough that no components of this switching current pulse falling in the range 200 to 400 mc are of sufficient amplitude to exceed the receiver noise level. If the use of such a switch were contemplated at lower frequencies, it would be necessary to use some form of balanced switching circuitry so that the current in the inductance of the filters would not change when the condition of the diodes was reversed. One arrangement which meets this requirement is shown in Figure 8. In this case, the two components of the switching current in the output transformer are in the same direction and hence cause no net induced voltage in the transformer secondary. However, the desired signal is fed in a "push-pull" connection so that the two components in the output transformer are of opposite phase and, therefore, produce a secondary voltage which is proportional to the sum of the two primary components. Careful attention to the construction of the transformer and to the matching of the diode characteristics should permit reduction of the unbalanced components of control pulse current to the order of 1% or less.

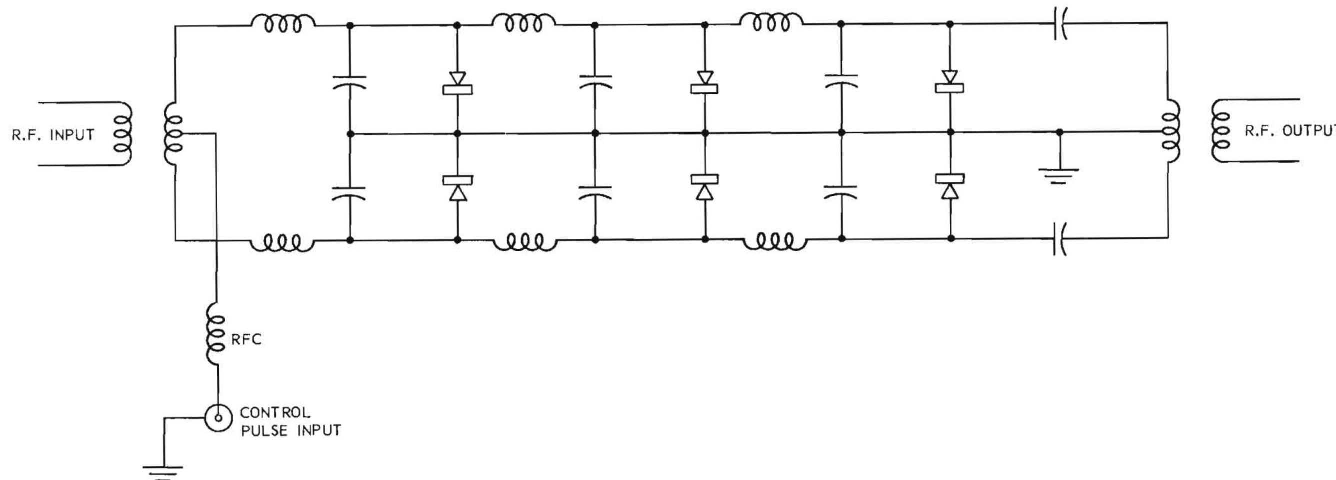



Figure 8. Balanced Switch

4.4 Equipment Design. The block diagram of Figure 9 illustrates the way in which the pulse control switch has been combined with the auxiliary receiver and suitable pulse delay and shaping circuitry to acquire proper synchronization of the pulse control switch for the suppression of periodic pulse interference. Referring to Figure 9, the desired signal and interference are superimposed at the

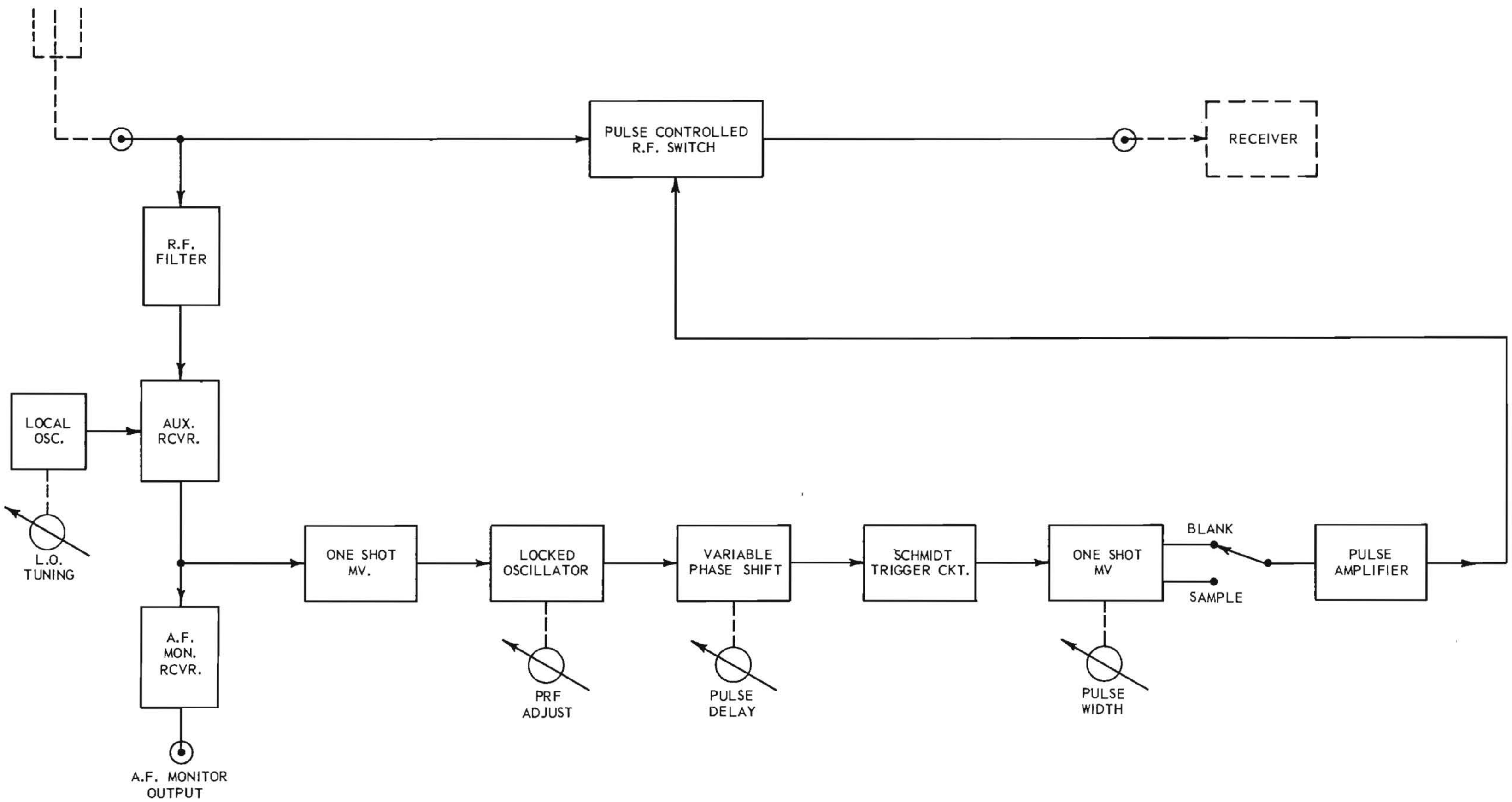



Figure 9. Block Diagram - Interference Blanker.

antenna and are applied both to the auxiliary receiver and to the pulse controlled switch. The auxiliary receiver supplies at its output one pulse for each pulse of the input interfering signal. Since the shape of this pulse depends on the shape of the transmitted interfering pulse, a one shot multivibrator is triggered by the output of the auxiliary receiver to produce a standard pulse shape which is independent of the shape of the input trigger pulse. This standard output pulse occurs at the same repetition rate as the input interfering pulse and is used to synchronize a locked oscillator at one of the harmonics of the repetition rate which lies in the frequency range 9 to 11 kc. The reason for this synchronization on one of the harmonics of the interfering repetition rate, rather than on the fundamental, follows. When the incoming desired signal is turned on and off by the blanking of the RF switch, a "hole" is created in the envelope of the desired signal. This "hole" constitutes an amplitude modulation of the desired signal at the blanking pulse rate. Since the repetition rate of most interfering pulse sources lies in the audio range, blanking the signal at the fundamental rate of the interfering pulse source would cause a corresponding amplitude modulation of the desired signal at an audio rate and this would result in an annoying tone interference in the output of the receiver; however, if the blanking pulse rate is at some harmonic of the input interfering pulse rate, the amplitude modulation of the desired signal will still occur but the lowest frequency component will be at the rate at which the blanking is being accomplished. For the case of the locked oscillator in question, this rate would be in the range 9 to 11 kc. As a result, low-pass filtering in the audio output of the receiver is effective in removing this tone, but does not affect the desired audio signal.

The sinusoidal output of the synchronized oscillator then is supplied through a continuously variable phase shifter to actuate the Schmidt trigger circuit which produces an output pulse at the zero crossing of each of the cycles of the phase shifter output. A variation of the phase shift causes a corresponding variation in the zero crossings of the phase shifter output and a resulting shift in the time position of the output pulse of the Schmidt trigger circuit. This output pulse is used to trigger a one shot multivibrator whose output pulse width is controllable. This output pulse, of the proper width to blank out the interfering signal, is passed through a pulse amplifier to control the RF switch in series with the input to the receiver. In operation, then, the interfering pulse is picked up by the auxiliary receiver whose output is used to synchronize pulses in the frequency range 9 to 11 kc, whose width and time of occurrence are manually adjustable. These pulses are used in the proper polarity to gate the input to the receiver off or on as the circumstances may require. A more detailed discussion of the individual components of this interference blanker is given in the following paragraphs.

4.4.1 Auxiliary Receiver. The requirements of an auxiliary receiver to detect the presence of interfering pulses and supply synchronization information to the locked oscillator depend upon the expected levels of the interfering pulse signals. For those situations where only large amplitude pulse interference will be encountered, a simple crystal-video receiver has sufficient sensitivity. A typical receiver of this type is illustrated in Figure 10. The principal limitation on the sensitivity of receivers of this type is the lack of diode nonlinearity in the region near zero bias. Although this can be remedied to some extent by the application of a small forward bias to the diode, generally, rectification efficiency at signal levels below several millivolts is quite poor. It is necessary to provide some means of RF selectivity in front of the diode detector in order that pulse signals whose frequencies are far removed from the frequency range in

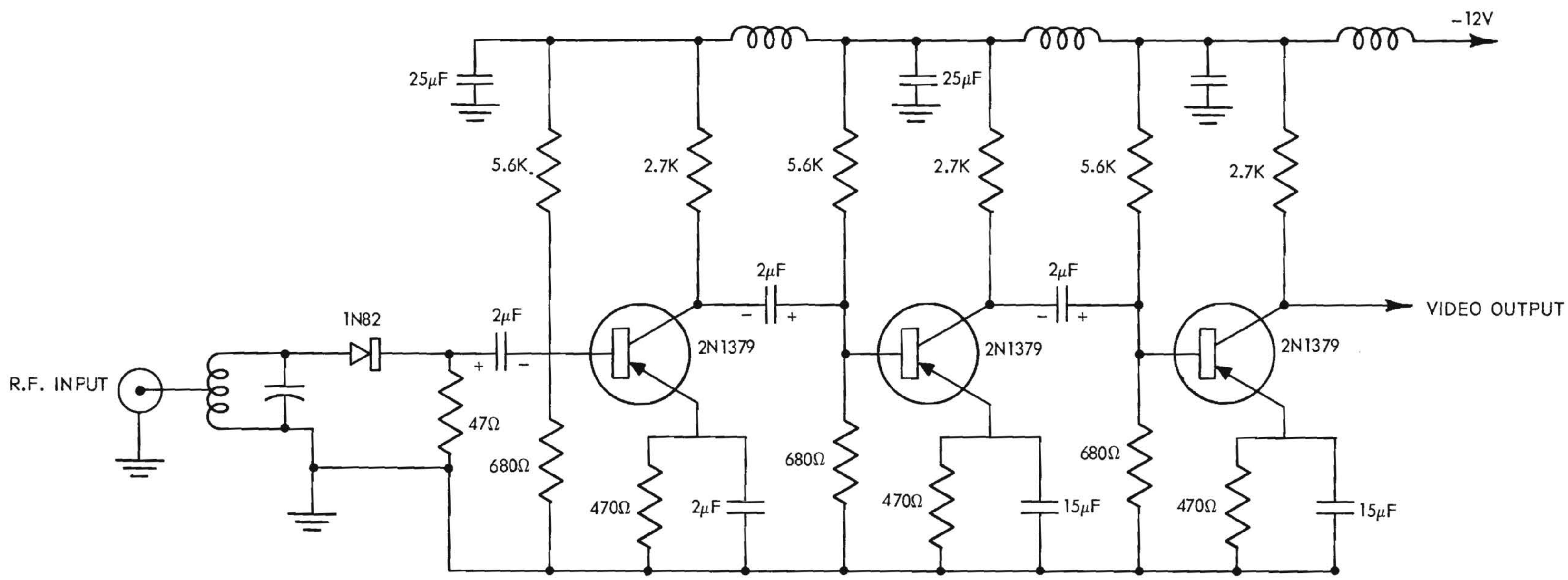



Figure 10. Crystal-Video Receiver.

which operation is contemplated do not produce false triggering of the blanking circuitry. Since no stringent selectivity requirement need be placed on this filter, adequate protection against this false triggering may be acquired with a simple single tuned RF filter. One such filter which has been successfully applied to this application is shown in Figure 11. This filter has essentially 50-ohm input and output impedances and has an effective "Q" of several hundred. The short section of two-wire transmission line is effectively short circuited by the ground plane to which it is attached. Since the length of this transmission line is considerably less than one-quarter wave length over the frequency range 200 to 400 mc, the reactance presented at the open end of this line section is essentially inductive. This inductance is resonated by a variable capacitor placed directly across the open end of the transmission line. One major disadvantage in the use of this particular configuration lies in the fact that both the rotor and stator of the variable capacitor are at RF potential with respect to the ground plane. This difficulty was overcome by mounting the capacitor in a polystyrene block and mechanically coupling to the rotor with an insulated shaft. The entire assembly was then enclosed in a shield can with adjustment of the resonant frequency being made by means of an insulated shaft. The input and output connections are made by means of coaxial cables which connect to the BNC connectors on either side.

Listening tests have demonstrated that when pulse signals of one-half millivolt or larger are superimposed on desired signals of the level of 2 to 4 microvolts, an appreciable interference occurs. For this reason, it is necessary to provide blanking action at the input to the receiver whenever the level of interfering pulses exceeds one-half millivolt. Since adequate sensitivity could not be obtained with a simple crystal-video receiver to operate at these low signal levels, it was necessary to construct a superheterodyne receiver to acquire the desired sensitivity.

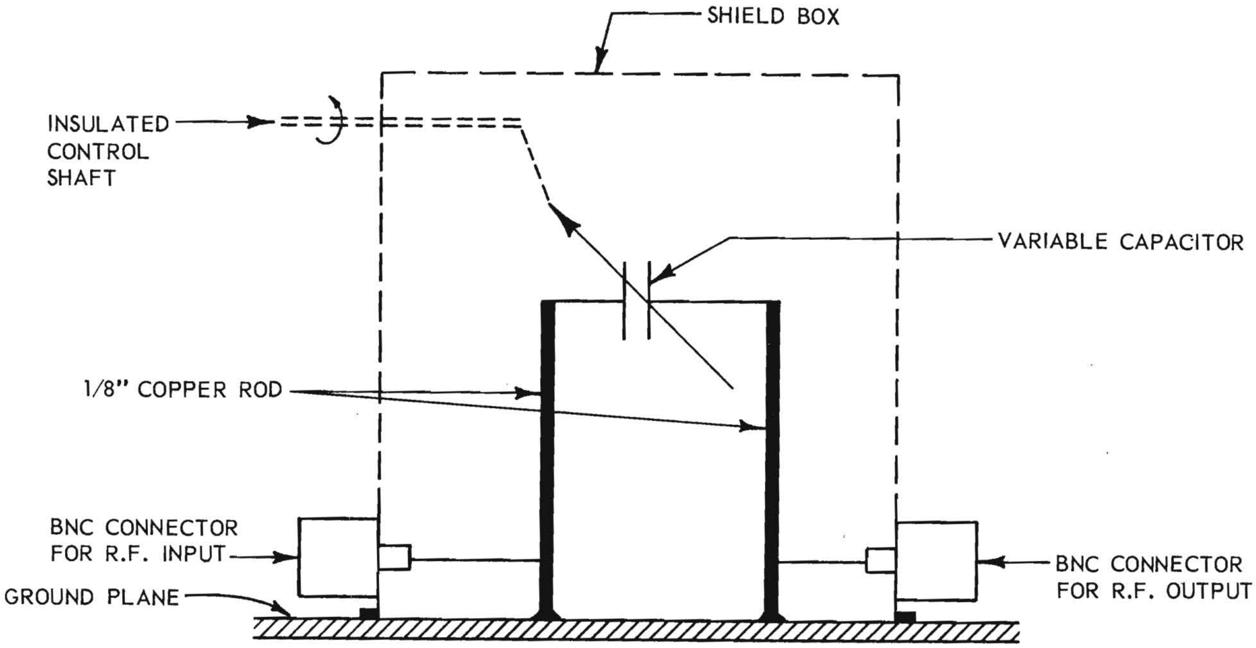



Figure 11. Tunable RF Filter

The physical construction of the receiver is illustrated in the photograph of Figure 12 and a schematic is given in the Appendix. A simple crystal mixer is used to heterodyne the output of the RF filter with the local oscillator to produce a 40 mc IF signal. The 40 mc IF frequency was chosen to obtain the necessary bandwidth for undistorted reproduction of the interference pulse. Three stages of IF amplification are provided to raise the signal level to a value sufficiently large to obtain efficient rectification in the diode detector. The output pulse from the detector is then used to drive a three-stage video amplifier which raises the level of the detected pulse to a value sufficiently large to insure reliable triggering of the one-shot multivibrator. An additional one-stage audio amplifier is connected to the video output to provide an audio monitoring output for use as an aid in receiver tuning. In order to simplify the

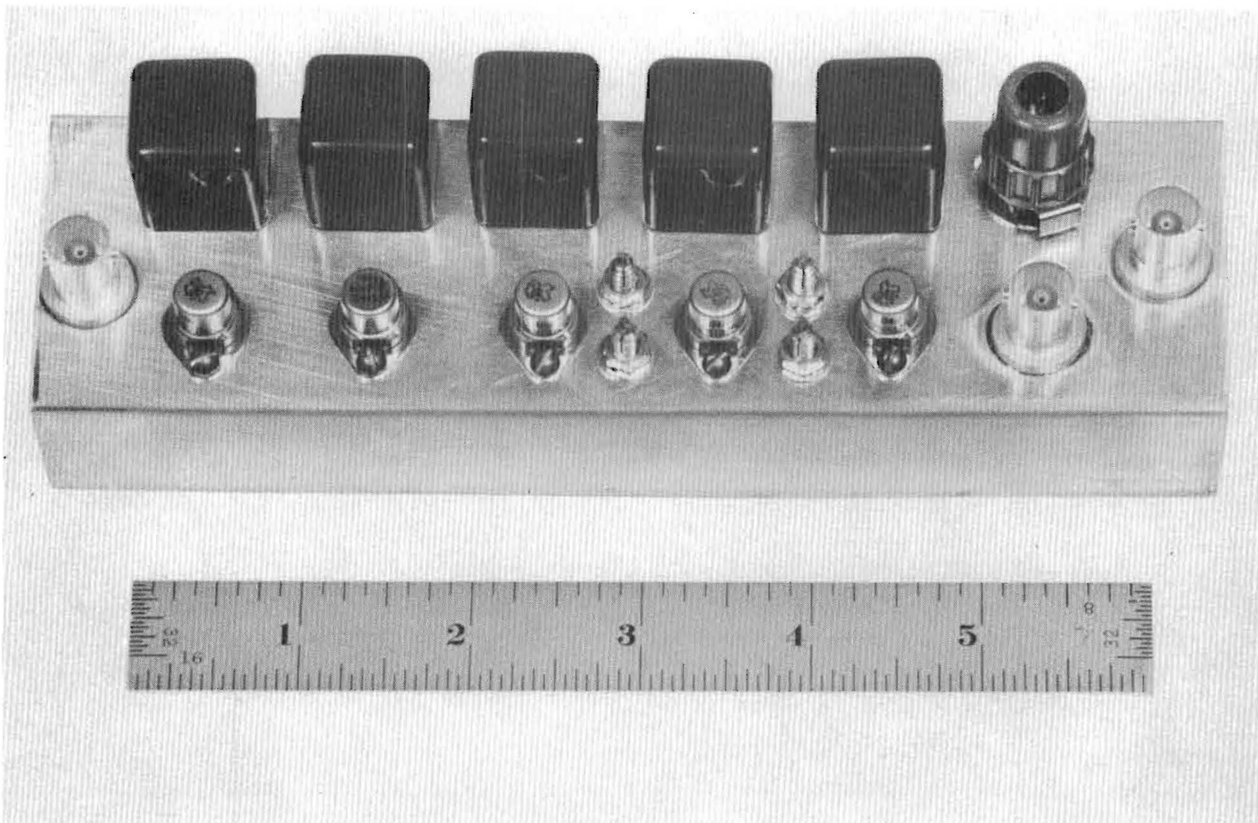



Figure 12. Auxiliary Receiver

receiver construction and to avoid tracking problems, the RF filter and the locked oscillator are separately tuned; the proper setting being obtained by maximizing the output of the audio monitoring amplifier on the interfering pulse signal.

4.4.2. Locked Oscillator. Figure 13 illustrates schematically the construction of the locked oscillator, along with the one-shot multivibrator which is used to standardize the synchronizing pulse for the locked oscillator. The

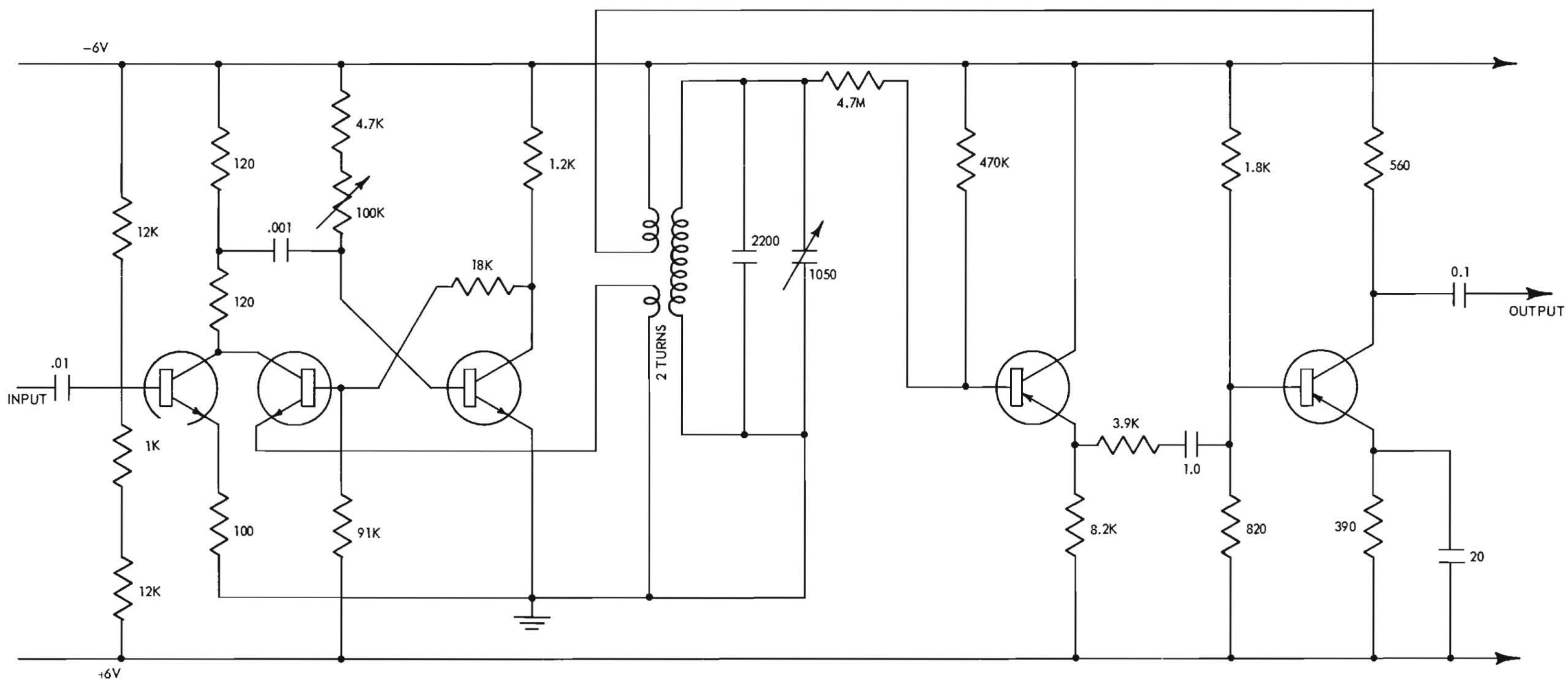



Figure 13. Locked Oscillator.

first transistor in the multivibrator is used simply as a means to parallel trigger the last two transistors, which form the multivibrator proper. Timing of the output pulse width is provided by proper selection of the coupling capacitor and is set for approximately 35 microsecond pulse width. The output pulse is taken as a current sample in the emitter of one of the transistors in the multivibrator. This current pulse is passed through a one-turn winding on the resonant circuit which controls the frequency of the locked oscillator. Due to the high "Q" of the oscillator coil, no appreciable circulating current flows in the oscillator tank circuit at frequencies only slightly removed from the resonant frequency of this circuit. Consequently, when the natural frequency of the oscillator is made nearly equal to one of the harmonic frequencies contained in the current pulse from the multivibrator, the only frequency component acting in the oscillator tank circuit to produce synchronization of the oscillator is that component which exists in the neighborhood of the resonant frequency of the oscillator circuit. For an interfering pulse rate in the neighborhood of 1,000 cycles, which is a typical value, the tenth harmonic of the pulse in the output of the multivibrator will produce across the tuning capacitor of the oscillator a voltage which is approximately 20 per cent of the voltage existing at this point due to the natural oscillation of the circuit. Since the "Q" of the oscillator coil is approximately 100, the frequency range over which synchronization can be effected may be calculated by the following relation:

$$BW_{(sync)} = \frac{\omega_o}{2Q_o}\left(\frac{V_s}{V_o}\right) \tag{2}$$

where
$$\begin{cases} \omega_o = \text{natural frequency of oscillation} \\ Q_o = \text{Q of oscillator resonant circuit} \\ V_s = \text{amplitude of the synchronizing signal} \\ V_o = \text{amplitude of the natural oscillation} \end{cases}$$

From this relationship, it is seen that a synchronizing bandwidth for the tenth harmonic of approximately 60 cycles is obtained. This is sufficiently wide to accommodate small drifts in the repetition rate of the interfering pulse source without the need for manual readjustment of the oscillator frequency. As a practical matter, tests were made on the synchronizing action of the oscillator in the range 9 to 11 kc when the repetition rate of the interfering pulse signal was varied. It was found that satisfactory synchronization could be maintained when the repetition rate was as low as 50 cps. The oscillator proper consists of a two-stage transistor feedback amplifier in which the feedback is coupled to the tuned circuit by means of a small two-turn inductive link on the oscillator tank circuit. The input of the transistor amplifier is isolated from the tuned circuit by means of a 4.7 megohm resistor which serves to make impedance variations at the input of the amplifier have negligible effect on the tuned circuit. In addition, it permits the maintenance of a low amplitude of natural oscillation in the tuned circuit so as to provide the maximum synchronization bandwidth. The output is taken from the collector of the output amplifier in the oscillator circuit.

4.4.3. Variable Phase Shifter. The output of the locked oscillator is supplied to a three-stage variable phase shifter which positions the point of zero crossing of the sine wave output of the locked oscillator so the blanking pulses which are generated at these points of zero crossing may have their time position of occurrence varied. The construction of the phase shifter as illustrated in the schematic in the Appendix is a familiar RC network whose phase shift is variable from 0 to $180^{\circ}$ but whose gain is independent of frequency. Since the time position of the output pulses must be varied over a range of almost one full period, it is necessary to provide $360^{\circ}$ of phase shift to insure that the range is adequately covered. Since the 0 and $180^{\circ}$ points in the phase shifter occur for the extreme values of RC = 0 and RC = $\infty$, it is necessary to provide three sections of phase shift to insure adequate coverage of the desired range without the necessity of operating at either of the two extremes. In the construction of the phase shifter, the phase in each section is controlled by one gang of a three-gang potentiometer so that a single control will give simultaneous variation of the phase shift in all three sections. Biasing for the three split loaded phase inverters is provided by a return of the base bias resistors to the collectors. This connection provides a degree of DC stability due to the negative feedback while at the same time maintaining a relatively high input impedance due to the large value of base biasing resistor used. The unbypassed emitter circuit also serves to raise the input impedance. Though this connection does not represent a necessarily optimum connection for obtaining high input impedance, it does supply the necessary unloading of the phase shift network and simultaneously gives the required phase inversion for driving the next succeeding phase shift network. Emitter followers are used at both the input and the output of the phase shifter in order to provide isolation of these terminals from external circuitry.

The output of the phase shifter is used to drive a Schmidt trigger circuit which fires at the zero crossing of the output of the phase shifter. This Schmidt trigger circuit in turn triggers a one-shot multivibrator; the proper trigger shape for this operation being obtained by means of a small differentiating transformer connected in the output of the Schmidt trigger. A diode is placed across the secondary of this transformer to remove the positive going portion of the wave form which might otherwise cause difficulty in triggering the one shot multivibrator.

The one shot multivibrator pulse width is adjustable from 5 to 35 microseconds so that variable width blanking pulses or sampling pulses may be generated. These output pulses are used to drive a pulse amplifier whose output stage is a power transistor capable of supplying the necessary 400 milliampere pulse to actuate the RF switch. Operation in either the blanking or the sampling mode is made possible by the simple expedient of reversing the connection of the output of the one shot multivibrator from one collector to the other.

4.4.4. Equipment Adjustment. The completed equipment is pictured in the photographs of Figures 14 and 15, which show the internal construction and parts layout as well as arrangement of controls on the front panel. With the exception of the auxiliary receiver and its local oscillator, the circuitry is constructed on phenolic boards which are arranged for simple removal from a central mounting structure in the center of the chassis. Each board is equipped with an easily removable plug which contains all connections to that particular board.

In order to place the equipment into operation, the antenna connection to the receiver is removed and instead is connected to the antenna output jack in the back of the chassis of the interference blanker. An output cable is then run from the output jack at the rear of the interference blanking chassis to the input jack of the receiver. Using the audio monitor output as an indication of proper tuning, the auxiliary receiver is tuned to the interfering pulse signal. Then, with the blanking pulse width set to its widest position, the PRF control which tunes the locked oscillator is varied until synchronization of the locked oscillator with the interfering pulse signal is obtained. Then by narrowing the

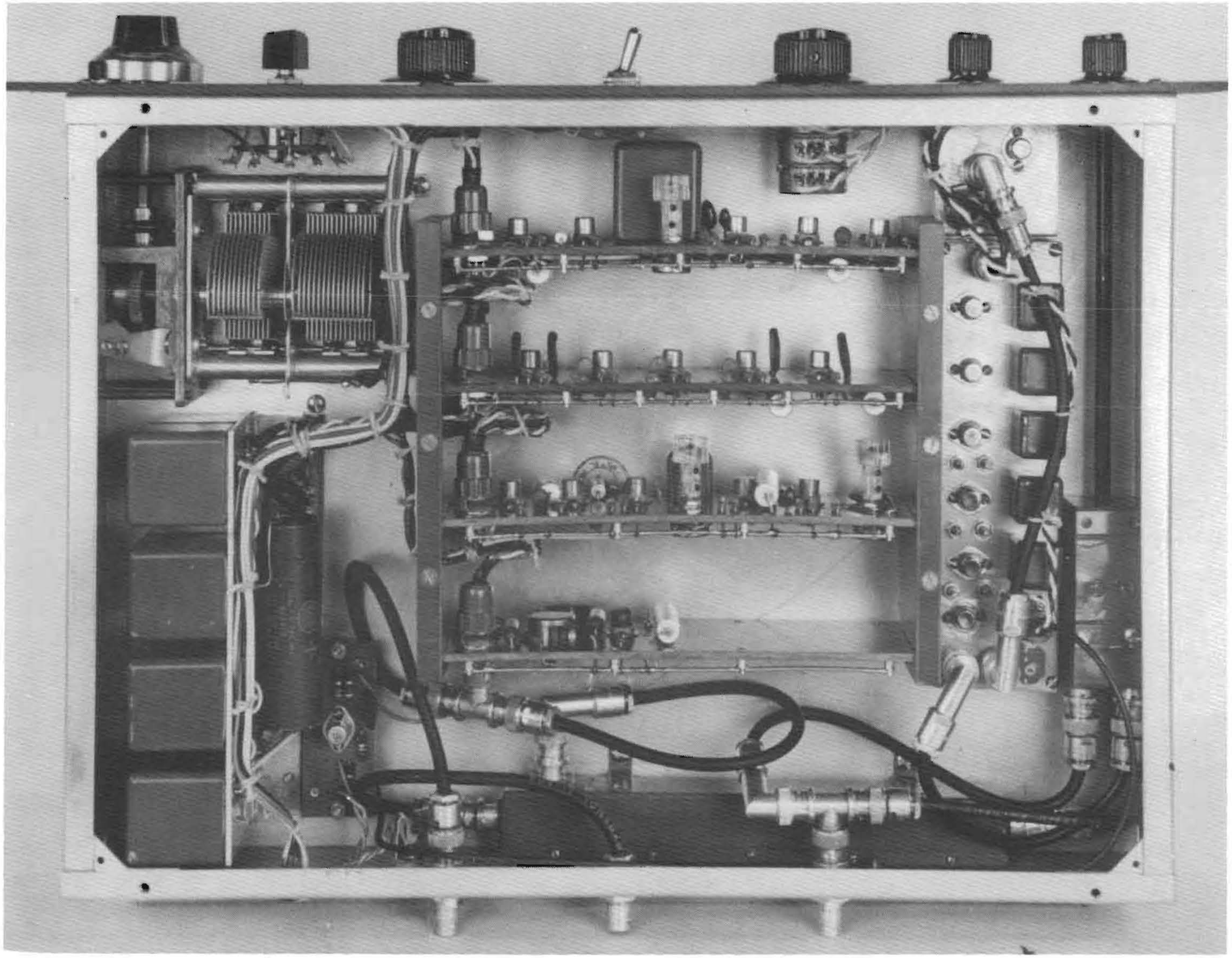


Figure 14. Internal Construction - Interference Blanker.

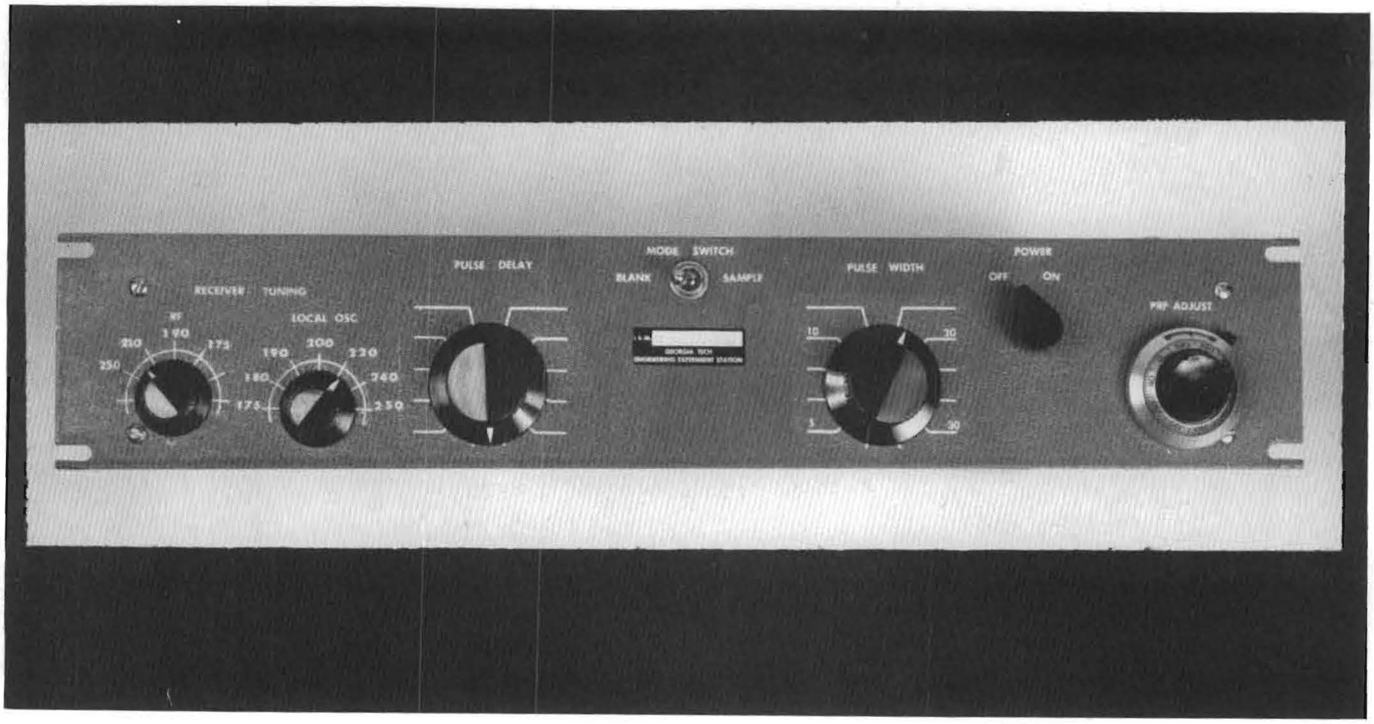



Figure 15. Front Panel - Interference Blanker

width of the blanking pulse and by adjustment of the pulse delay, proper position of the blanking pulse with respect to interfering pulse may be obtained. In this manner, interfering signals may be easily removed from the input to the receiver and excellent reproduction of desired signals at levels of a few microvolts can be obtained.

## 5. ACKNOWLEDGEMENTS

This program was conducted under the general supervision of W. B. Wrigley, who is head of the Communications Branch. P. A. Kresge also made significant contributions to the work described in this report.

Respectfully submitted:

W. B. Warren, Jr.
Project Director

Approved:

M. W. Long, Chief
Electronics Division

# 6. APPENDIX

6.1. Analysis of the Sampling Technique. The following analysis points out the relationship between the input and reconstructed AM signals in more detail. Referring to Figure 3, the output of the sampling switch is given by

$$E_o(t) = E_i(t) \quad T(t), \tag{3}$$

Where $E_i(t)$ is the input signal to the switching device and $T(t)$ is the switch transfer function. Since the switch is always on or off, the transfer function $T(t)$ must be of a form similar to Figure 16.

Here the switch is operating periodically at the frequency $f_s$. Because of this periodicity, $T(t)$ may be expanded in a Fourier series,

$$T(t) = \frac{A_0}{2} + \sum_{n=1}^{\infty} A_n \cos (n2\pi f_s t), \tag{4}$$

which, upon substitution in (1), gives

$$E_o(t) = E_i(t) \left\{ \frac{A_0}{2} + \sum_{n=1}^{\infty} A_n \cos (n2\pi f_s t) \right\}. \tag{5}$$

If $E_i(t)$ is a narrow band signal such as the AM signal of Equation (6),

$$E_i(t) = \left[ 1 + m \cos (2\pi f_m t) \right] \cos 2\pi f_o t. \tag{6}$$

$$\text{Then } E_o(t) = \left\{ \left[ 1 + m \cos (2\pi f_m t) \right] \cos 2\pi f_o t \right\} \left\{ \frac{A_0}{2} + \sum_{n=1}^{\infty} A_n \cos (2\pi n f_s t) \right\} \tag{7}$$

or

$$E_o(t) = \frac{A_0}{2} E_i(t) + \sum_{n=1}^{\infty} \frac{A_n}{2} \cos (2\pi) (f_o \pm nf_s)t$$

$$+ \sum_{n=1}^{\infty} \frac{mA_n}{4} \cos (2\pi) (f_o + f_m \pm nf_s)t \tag{8}$$

$$+ \sum_{n=1}^{\infty} \frac{mA_n}{4} \cos (2\pi) (f_o - f_m \pm nf_s)t.$$

The band-pass filter removes all except the first term so that the output signal is

$$E_1(t) = \frac{A_0}{2} E_i(t). \tag{9}$$

From Equation (9), it is seen that the desired signal gain is

$$\text{Gain} = \frac{A_0}{2} \tag{10}$$

which is simply the average value of the transfer function of the switch. For the rectangular pulses of Figure 16,

$$\frac{A_0}{2} = \text{average value of } T(t) = \frac{\Delta}{T} = \text{duty cycle of the switching signal.} \tag{11}$$

6.2. Spurious Responses. In sampling or blanking the front end of a receiver in order to reduce the effects of pulse interference, several undesirable responses arise which are due to the switching process. These responses can be eliminated only if sufficient selectivity is placed ahead of the switching element to reject signals whose frequencies lie on one of these response frequencies.

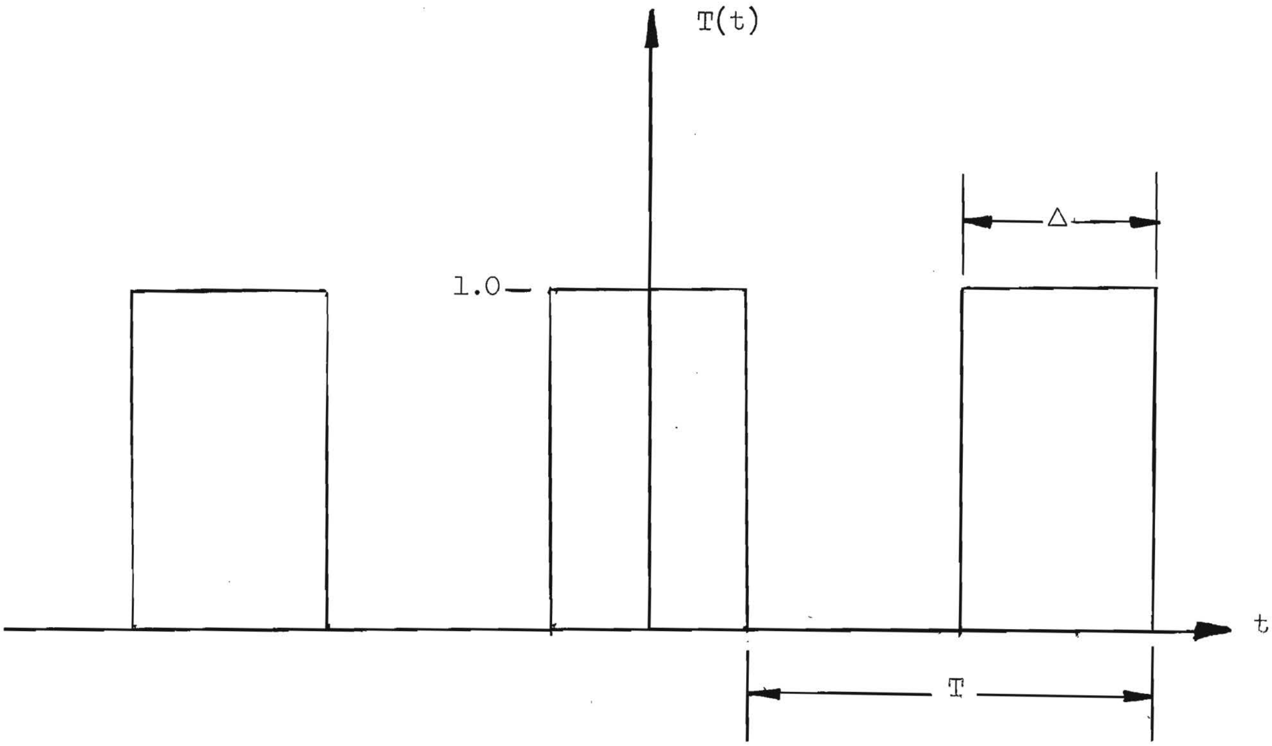



Figure 16. Sampling Switch Transfer Function

These responses arise because of the switching at the blanking or sampling rate. The following analysis points out the location and relative magnitudes of these responses:

If, referring to Equation (5),

$E_i(t)$ is a single tone signal of frequency $f_i$, i.e. (12)

$E_i(t) = B \cos (2\pi f_i t),$

then

$$E_o(t) = B \cos (2\pi f_i t) \left\{ \frac{A_0}{2} + \sum_{n=1}^{\infty} A_n \cos (2\pi n f_s t) \right\} \tag{13}$$

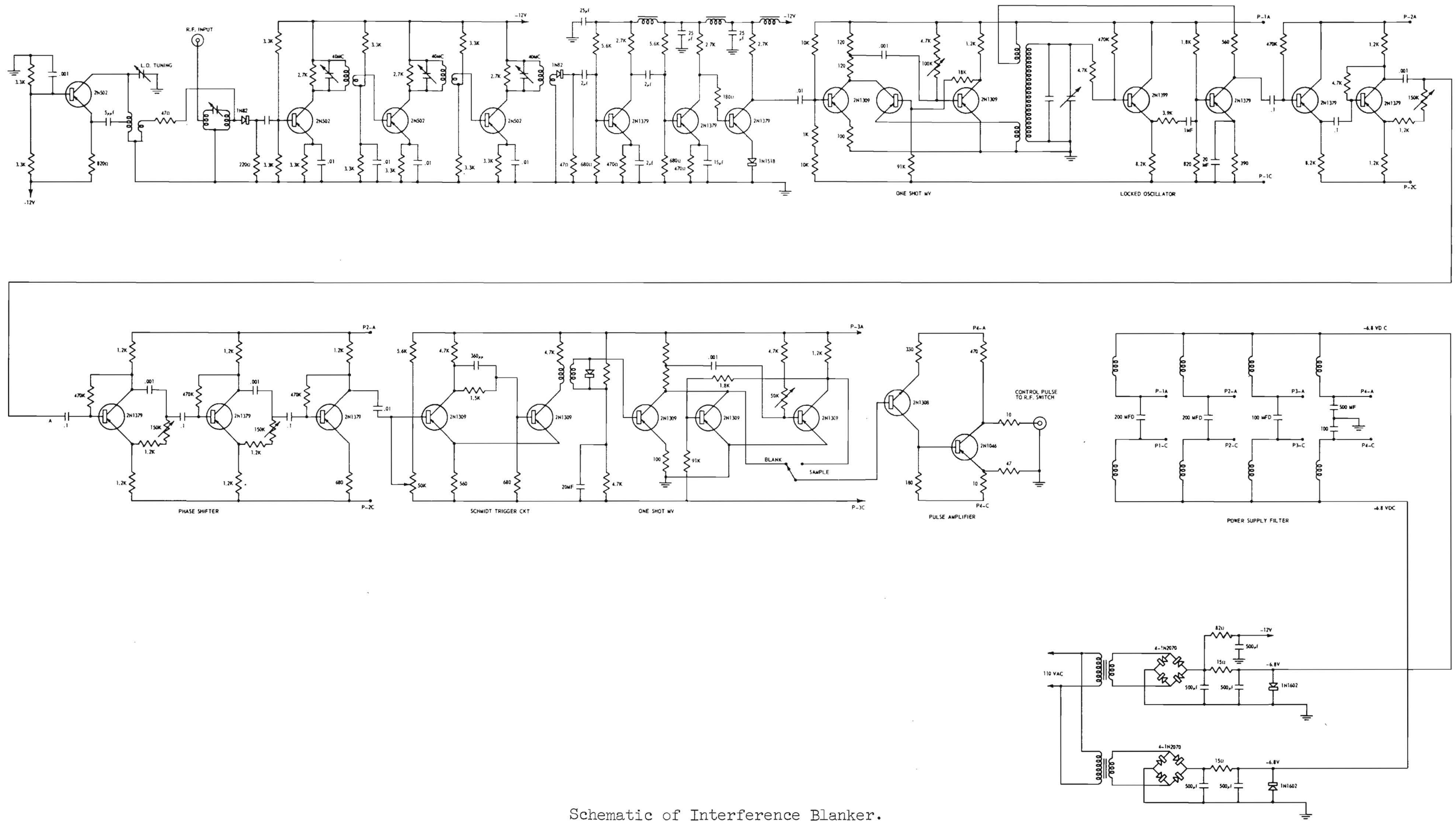
or

$$E_o(t) = \frac{A_0}{2} B \cos (2\pi f_i t) + \sum_{n=1}^{\infty} \frac{A_n B}{2} \cos 2\pi (f_i \pm n f_s) t. \tag{14}$$

Now the frequencies, $f_i$, are determined so that $E_o(t)$ has the components at the tuned frequency of the receiver, $f_o$. There are two conditions which satisfy this requirement. They are:

(1) $f_i = f_o$ — This is the output of the switching device due to a signal at the operating frequency, $f_o$.

(2) $f_i = f_o \pm n f_s$ — These values of $f_i$ give the frequency positions of the spurious responses, the relative amplitudes being determined by the Fourier Coefficients, $A_n$, in Equation (4).




Schematic of Interference Blanker.

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# COMMUNICATIONS INTERFERENCE REDUCTION STUDY

by W. B. Warren, Jr.

FINAL REPORT

Project No. A-525

Contract No. AF 30(602)-2399

Prepared for
Rome Air Development Center
Air Force Systems Command
United States Air Force
Griffiss Air Force Base
New York

3 December 1962

Engineering Experiment Station
GEORGIA INSTITUTE OF TECHNOLOGY
Atlanta, Georgia

**Title of Report**

RADC-TDR-62-631

**PUBLICATION REVIEW**

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Directorate of Communications

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RADC-TDR-62-631

# COMMUNICATIONS INTERFERENCE REDUCTION STUDY

By

W. B. Warren, Jr.

ENGINEERING EXPERIMENT STATION
of the Georgia Institute of Technology
Atlanta, Georgia

FINAL REPORT

PROJECT NO. A-525

Contract No. AF 30(602)-2366

Prepared for

Rome Air Development Center
Air Force Systems Command
United States Air Force
Griffiss Air Force Base, New York

## FOREWORD

This is the Final Technical Report covering work performed under Contract AF 30(602)2366 and deals with the investigation of techniques to reduce interference to communication and measurements equipment.

One technical note was published during the course of this project. This note deals with the design and construction of a device to suppress high level pulse interference to communications receivers. This program was conducted initially under the supervision of W. B. Wrigley and more recently under the direction of D. W. Robertson, who is Head of the Communications Branch. In addition to the author, the following staff members also participated in work described in this report: P.A. Kresge, D. G. Hobbs, C. S. Wilson, and J. Caudell.

Respectfully Submitted,

W. B. Warren, Jr.
Project Director

Approved:

M. W. Long, Chief
Electronics Division

## ABSTRACT

This report describes a collection of techniques which are effective in reducing or eliminating the effects of pulse and CW interference to communications and measurements receivers. The techniques covered include (1) blankers, (2) RF limiters, (3) cavity preselectors, (4) selective RF preamplifiers, (5) linear mixers, (6) tunnel diode mixers, (7) special audio filters, and (8) modifications to the R361 communication receiver.

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LIST OF FIGURES (Cont.)

## 1. Introduction

The recent increases in both the power and number of devices radiating energy in the UHF spectrum create an environment in which many currently used military communications receivers were not designed to operate. The use of these receivers in such an environment has in many instances resulted in annoying or impossible interference situations. In particular, the co-location of high-powered pulse modulated transmitters with narrowband communications receivers has produced serious overloading and desensitization of these receivers so as to render impossible the reception of weak, desired signals. In addition, the need for accurate data for use in the efficient prediction of possible interference situations requires an interference rejection capability that many currently used measurement receivers do not possess.

As a result, there has arisen a need for the development of new techniques and devices for the suppression of interference to communications and measurement receivers The work under this contract has been concerned with satisfying this need in certain areas. Specifically, this report presents a collection of techniques which have been developed to permit the successful operation of receivers in the presence of high level interfering signals. Whenever possible, a piece of working equipment has been constructed to demonstrate the feasibility of each technique presented. In most cases, sufficient information is given to enable others to construct similar equipments. An attempt has been made to design the equipment in such a way as to minimize the number of modifications that must be made in the receiver which a particular device or technique is intended to protect. In most instances, no internal receiver modifications are required in the application of these suppression devices.

## 2. Discussion

### 2.1 Receiver Gating Techniques for Pulse Interference

In many instances, it is necessary to co-locate narrow band communications equipment with high-powered, pulse-type equipment, such as radar. In this situation, it is not always possible to prevent a considerable amount of pulse energy from appearing at the input to the communications equipment. In some instances, this power level may be as high as several watts. Such power levels, incident upon conventional communications equipment, will cause serious overloading and desensitization and render impossible the reception of weak desired communications signals. Overloading of the receiver input stages by a pulse causes grid current to be drawn with a consequent rapid charging of the AGC line and severe desensitization of the receiver because of the excessive AGC voltage which is developed. Consequently, adequate suppression of the effects of high level pulse interference can be obtained only if the interfering pulse is rejected directly at the input terminals of the receiver.

In any interference suppression device, it is necessary to recognize some essential difference between the desired and interfering signals and to make use of this difference to obtain the desired interference reduction. In the case of a narrow band desired signal, which is modulated with speech or other narrow band intelligence and which is being interfered with by a pulse type signal, one essential difference lies in the fact that the interference is limited in time, while the desired signal is not. Thus, by switching the input to the narrow band receiver on or off at the proper instants of time, the interfering signal can be drastically reduced in amplitude or completely eliminated. On the other hand, the desired signal, not being of a time limited character, is sufficiently unaffected by this gating of the receiver input to permit the desired signal to be successfully recovered.

<u>2.1.1 Blanking</u>: This gating technique can be applied in either of two basic ways. The first of these is called "blanking". In the blanking technique, the receiver is turned off for the duration of each interfering pulse and turned on again between pulses, but the gaps in the desired signal due to this gating process do not seriously degrade the intelligibility of the desired signal. The action of the pulse controlled switch in removing the interfering pulse signal is indicated by the wave forms of Figure 1, which illustrate the situation of a desired sine wave modulated AM signal with a periodic pulse interference superimposed. Notice that this technique will be effective only if the blanking pulses are synchronous with the interfering pulse signal. In cases where it is possible, the best method to obtain this synchronization is to have a direct cable connection with the source of interference so that a blanking pretrigger will be available to compensate for the delay in the generation of the blanking pulse and in the disconnect action of the blanking switch. Unless some form of pretrigger is available, it is not possible to blank a particular interference pulse by detecting that particular pulse and using this detected output to generate the necessary control pulse for the blanking switch unless an adequate delay in the main signal path can be obtained. In general, such a delay in the signal path is difficult to obtain and has the additional disadvantage that most

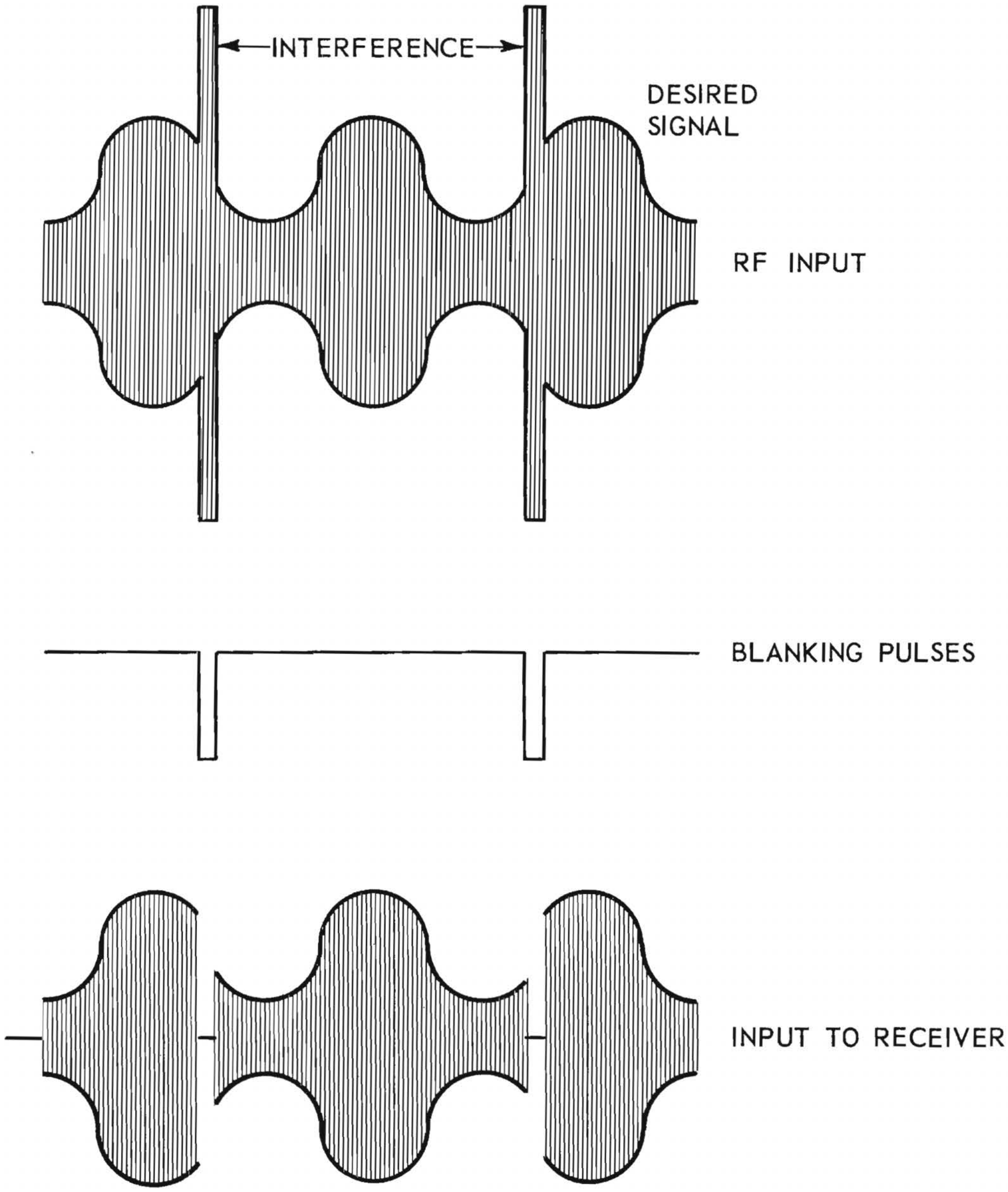



Figure 1. Blanking Waveforms

devices which might be used for this purpose exhibit an insertion loss that would unduly attenuate the desired signal. However, if the proper pretrigger pulse is available, the blanking switch may be operated slightly before the arrival of the interference pulse at the receiver input and complete blanking of the interfering signal can be obtained. If a direct cable connection to the source of pulse interference is not available, an auxiliary receiver can be used to detect the interfering pulses and supply them to a blanking pulse generator as synchronizing information. In this event, it is not possible to have the proper triggering information due to the delay which is associated with the switch and the auxiliary receiver. However, for periodic interference, the necessary triggering information can be obtained by delaying the interfering pulses by an amount just short of one pulse interval. In this arrangement, the blanking pulse corresponding to a particular interfering pulse is generated by the previous interfering pulse delayed by almost one pulse interval. This technique is effective as long as the condition of periodicity is met.

2.1.2 Sampling: A second receiver gating technique which may be used to reduce or eliminate pulse interference is called "sampling". In this process, samples of the input signal are taken at a rate which is sufficiently high to reconstruct all the essential information in the desired signal from knowledge of the samples alone. Since the required information in the input signal is the envelope (in the case of an AM signal), it is not necessary to provide samples at a rate high in respect to the carrier frequency of the input signal, but rather it is sufficient to provide samples at a rate which is high with respect to the highest frequency components contained in the envelope of the desired signal. For the case of a speech modulated signal occupying the frequency range 200 to 4,000 cps, the required minimum sampling rate is 8,000 per second, in accordance with well known theory of sampled signals. However, in a practical situation, the perfect filters required by the theory are not available and a slightly higher sampling rate is necessary to permit "clean" recovery of the original signal from knowledge of the samples alone. For the case mentioned above, a practical sampling rate of the order of 10,000 samples per second is necessary. Since the sampling of the input signal is done at the RF input to the receiver, it is necessary to reconstruct the desired signal from the sampled signal by means of a band-pass filter, rather than by a low-pass filter, as is usually the case.

Figure 2 illustrates the technique of sampling an amplitude modulated signal and reconstructing the original signal by a band-pass filter. Since the output of the sampling switch is zero, except at the sampling instants, it is apparent that the output of the band-pass filter is dependent only on the values of the input AM signal at these sampling times, since the input to this filter is zero at all other times. Hence, any interfering pulse signals which might occur at times other than the sampling instants will not be present in the output of the sampling switch and, therefore, will not be present in the output of the band-pass filter.

In a practical situation, the bandwidth of the RF amplifier portion of a conventional receiver is not sufficiently narrow to act as a band-pass filter, such as is assumed to be available in acquiring the wave forms of Figure 2. As a result, the input to the mixer and, hence, the input to the IF amplifier is in the nature of a pulse amplitude modulated signal, similar to that labeled "sampled output" in Figure 2, with the exception that the

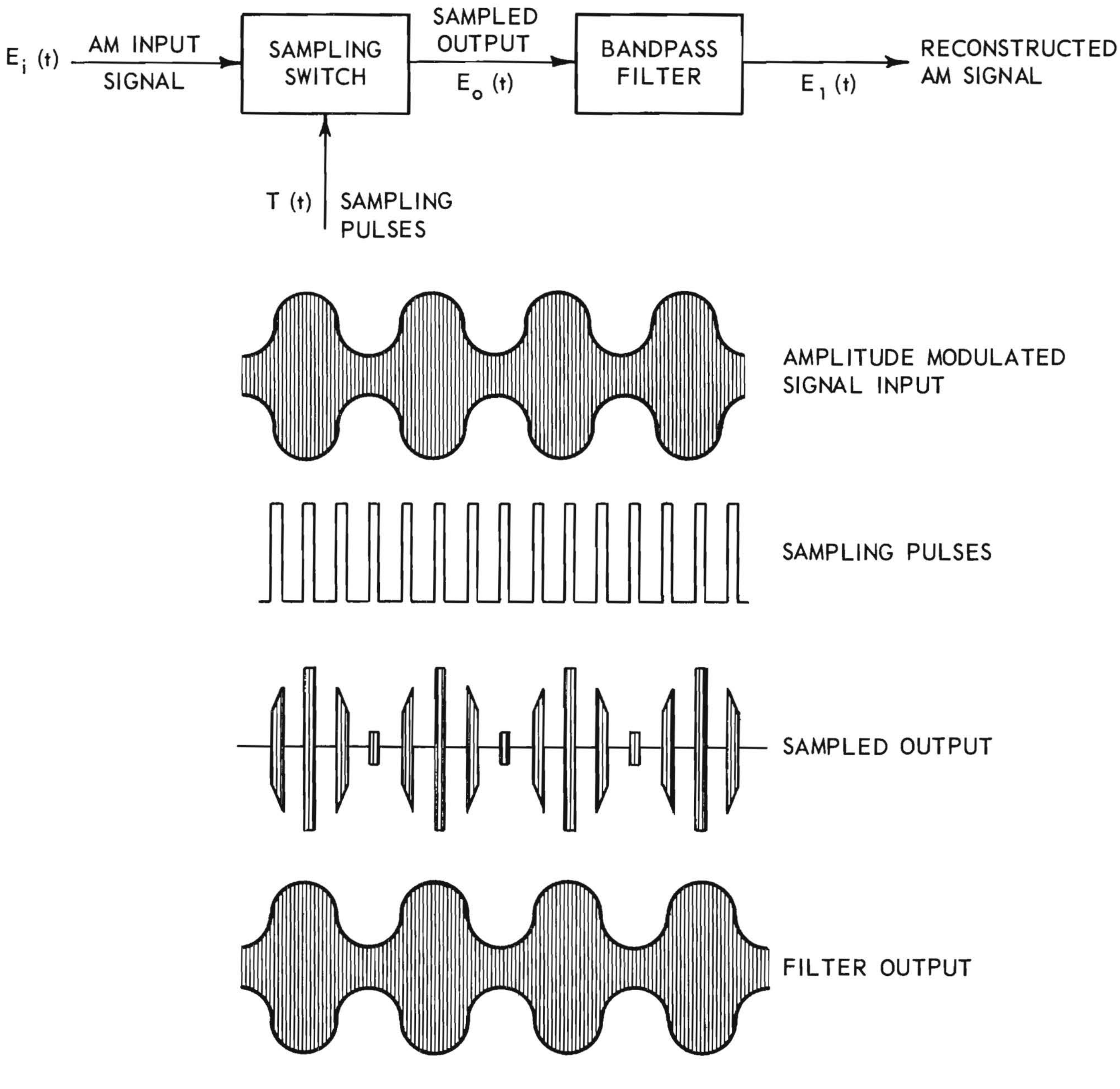



Figure 2. Sampling Waveforms

carrier frequency of this pulse amplitude modulated signal is now the IF frequency of the receiver, rather than that of the original input signal. The narrow bandwidth of the IF amplifier then serves as the required band-pass filter. The filter output will appear as the reconstructed AM signal and may be applied to an envelope detector to produce the desired envelope modulation at audio frequency.

The work of this project in developing equipment to use these principles in suppressing pulse interference is described in detail in "A Pulse Interference Blanker". This Technical Note was prepared under this contract and has the publication number RADC-TN-61-330. Consequently, only a general discussion of this equipment will be given here.

2.1.3 Diode Switch: In the application of either the blanking or sampling techniques to the reduction of pulse interference, it is necessary to have a switch which can be controlled by the blanking or sampling pulses to effect a rapid disconnection of the antenna so as to prevent interfering pulses from entering the receiver. At the same time, this switch must be capable of a very low minimum insertion loss when it is used to connect the receiver and the antenna together so as not to unduly attenuate low-level desired signals. The basic design of such a switch, capable of operation in the frequency range of 200 to 400 mc, is indicated in Figure 3. This switch is essentially a multi-section low-pass filter whose cut-off frequency is about 400 mc. Each section of this low-pass filter has a normally back-biased diode connection in shunt with it so that the effect of the diode is to add a small amount of shunt capacity to each section of the filter. This small added capacitance is taken into account in the design of the

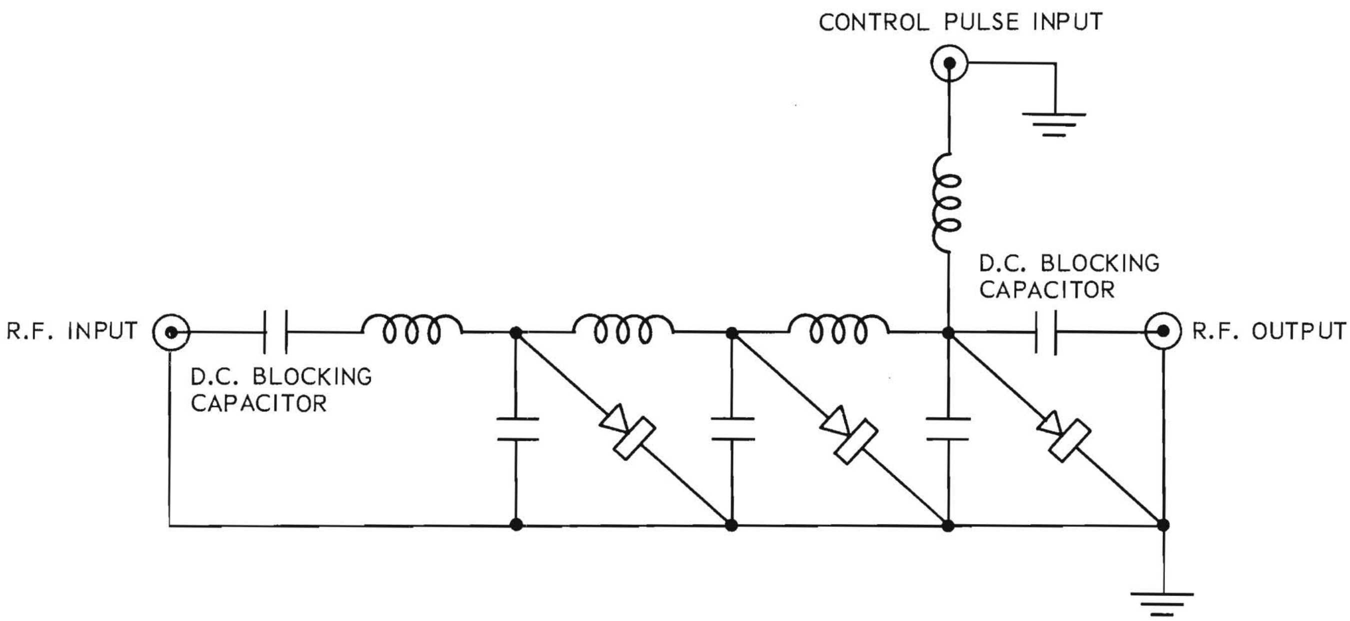



Figure 3. Diode Switch

filter so that the capacitive effect of the back-biased diode is removed and the attenuation in the pass band of the filter is quite small. However, the application of a forward biasing voltage to the shunt diodes causes them to exhibit a very low shunt resistance across each section of the filter, and the attenuation of the filter becomes very large. This large attenuation results from the considerable impedance mismatch encountered at the input to each section of the filter when the diodes are forward biased. As a result, only a small fraction of the power incident upon each section is transmitted by that section; the larger portion of the power being reflected back to the source.

Two capacitors are used at the input and output of the filter to isolate the filter for dc, but the reactance of these capacitors at frequencies in the 200 to 400 mc range is negligible. The control pulse is applied to the filter through an RF choke whose impedance is high with respect to the 50 ohm impedance level of the filter but does not present any appreciable reactance at frequencies contained in the control pulse.

A set of typical frequency characteristics of a diode controlled switch is shown in Figure 4. The insertion loss with a 400 milliampere control current exceeds 56 db over the range of 200 to 400 mc, while the insertion loss with the diodes back-biased is less than 1 db over the same frequency range. This 56 db insertion loss is sufficient to reduce a 10 watt interfering pulse appearing at the antenna terminals of the receiver to a level of $25 \times 10^{-6}$ watts. Although this level is quite low and will not cause serious overloading in the front end of the receiver, it is still sufficiently large to cause considerable difficulty in the reception of desired signals close to the sensitivity threshold of the receiver. It is also possible that this remaining level is large enough to produce a considerable AGC voltage at the output of the AGC detector, especially in those receivers having conventional peak detecting AGC rectifiers. This excessive AGC voltage may cause sufficient desensitization of the receiver to prevent the reception of the very weak desired signals. In such an event, two of the diode switches may be connected in cascade so that an additional 56 db of attenuation of interfering pulse signal is obtained. This amount of attenuation is sufficiently large to reduce the level of the interfering pulses almost to the noise level of the receiver. Even with this arrangement, the maximum insertion loss in the

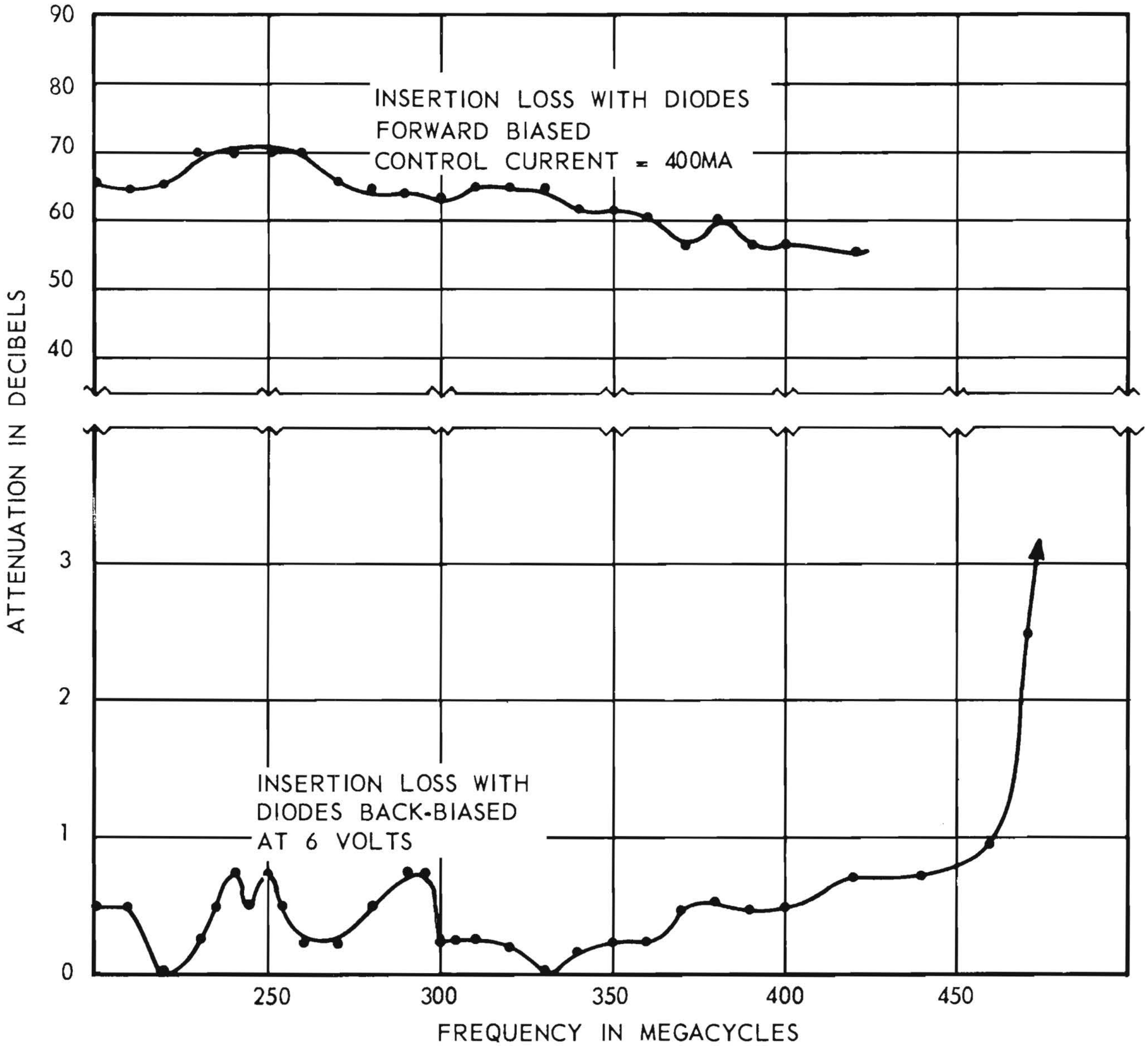



Figure 4. Switch Insertion Loss Characteristics

back-biased condition is considerably less than 2 db over the entire range of interest from 200 to 400 mc.

In the use of a switch of this nature, there is a change in the current in the inductances of the low-pass filter when the diodes are switched from the forward to the back-bias condition. Since a portion of this current flows in the output circuit of the switch, it is possible that extraneous signals may be presented to the input to the receiver even though no input signal is incident upon the antenna. Fortunately, this difficulty is not encountered in the use of this switch in the frequency range 200 to 400 mc, since the switching of the diodes from the forward to back-bias condition is slow enough that no components of this switching current pulse falling in the range 200 to 400 mc are of sufficient amplitude to exceed the receiver noise level.

2.1.4 Equipment Design: The block diagram of Figure 5 illustrates the way in which the pulse control switch has been combined with an auxiliary receiver and suitable pulse delay and shaping circuitry to acquire proper synchronization of the pulse control switch for the suppression of periodic pulse interference. Referring to Figure 5, the desired signal and interference are superimposed at the antenna and are applied both to the auxiliary receiver and to the pulse controlled switch. The auxiliary receiver supplies at its output one pulse for each pulse of the input interfering signal. Since the shape of this pulse depends on the shape of the transmitted interfering pulse, a one-shot multivibrator is triggered by the output of the auxiliary receiver to produce a standard pulse whose shape is independent of the shape of the input trigger pulse. This standard output pulse occurs at the same

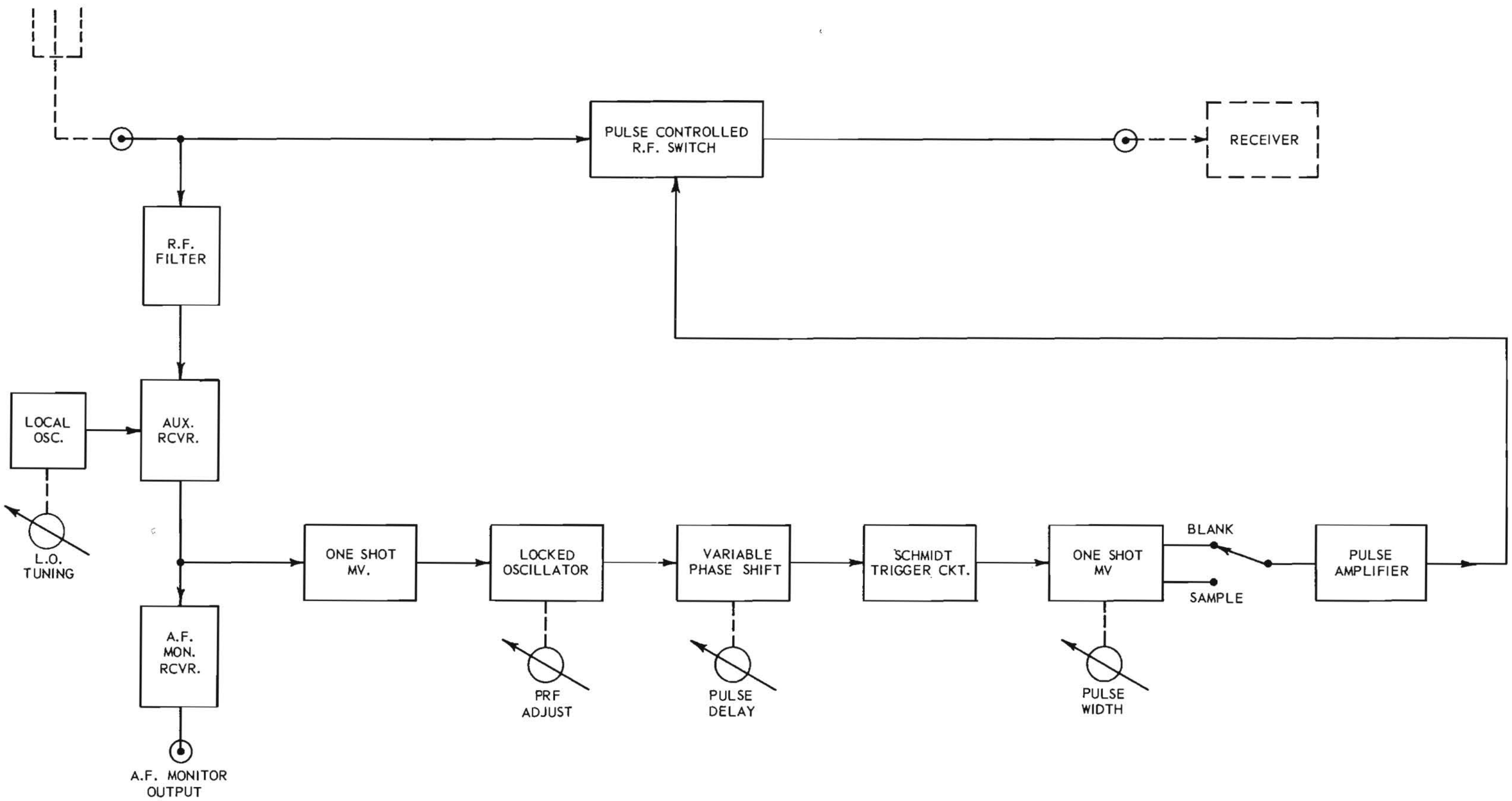



Figure 5. Block Diagram of Interference Blanker

repetition rate as the input interfering pulse and is used to synchronize a locked oscillator at one of the harmonics of the repetition rate which lies in the frequency range 9 to 11 kc. The reason for this synchronization on one of the harmonics of the interfering repetition rate, rather than on the fundamental can be explained as follows:

When the incoming desired signal is turned on and off by the blanking action of the RF switch, a "hole" is created in the envelope of the desired signal. This "hole" constitutes an amplitude modulation of the desired signal at the blanking pulse rate. Since the repetition rate of most interfering pulse sources lies in the audio range, blanking the signal at the fundamental rate of the interfering pulse source would cause a corresponding amplitude modulation of the desired signal at an audio rate and this would result in an annoying tone interference in the output of the receiver. However, if the blanking pulse rate is at some harmonic of the input interfering pulse rate, the amplitude modulation of the desired signal will still occur but the lowest frequency component will be at the rate at which the blanking is being accomplished. For the case of the locked oscillator in question, this rate would be in the range of 9 to 11 kc. As a result, low-pass filtering in the audio output of the receiver is effective in removing this tone, but does not affect the desired audio signal.

The sinusoidal output of the synchronized oscillator is supplied through a continuously variable phase shifter to actuate the Schmidt trigger circuit which produces an output pulse at the zero crossing of each of the cycles of the phase shifter output. A variation of the phase shift causes a corresponding variation in the zero crossings of the phase shifter output and a resulting shift in the time position of the output pulse of the Schmidt trigger circuit. This output pulse is used to trigger a one-shot multivibrator whose output pulse width is controllable. This output pulse, of the proper width to blank out the interfering signal, is passed through a pulse amplifier to control the RF switch in series with the input to the receiver. In operation, then, the interfering pulse is picked up by the auxiliary receiver whose output is used to synchronize pulses in the frequency range 9 to 11 kc, whose width and time of occurrence are manually adjustable. These pulses are used in the proper polarity to gate the input to the receiver off or on as the circumstances may require. In this manner, interfering signals may be easily removed from the input to the receiver and excellent reproduction of desired signals at levels of a few microvolts can be obtained.

## 2.2 RF Limiters

In those situations where the amplitude of an interfering pulse signal is large with respect to the amplitude of the desired signal, a considerable improvement in the signal to interference ratio may be obtained by preceeding the receiver with a limiting device to restrict the large amplitude excursions of the interfering signal. For such a limiter to be effective, its limiting threshold must be sufficiently low to restrict the range of interfering signal amplitude to values considerably below the overload level of the receiver. Several approaches were investigated in an attempt to provide a device with the proper limiting characteristics.

### 2.2.1 Non-linear Capacitor in Parallel with a Resonant Circuit:

The sketch of Figure 6 indicates a method in which a voltage sensitive capacitor was connected in parallel with a resonant circuit. In this arrangement, the positive peaks of the signal appearing across the tuned circuit cause the diode to conduct, charging the fixed capacitor to a voltage equal to the peak of the applied signal. When the signal swings in the negative direction, the diode is back-biased and exhibits capacitance which is a function of the dc voltage across the fixed capacitance. Since this dc voltage is equal to the peak of the input signal, the diode capacitance will vary with the amplitude of the applied signal. If the resonant circuit is initially tuned to resonance for small signals, then at large signal levels the bias voltage developed across the fixed capacitance will cause the resonant circuit to be de-tuned, and the gain to the applied signal will be reduced.

When a large amplitude signal was applied to the circuit of Figure 6, the desired limiting action was present when the frequency of the applied signal was below the resonant frequency of the circuit. However, when the frequency

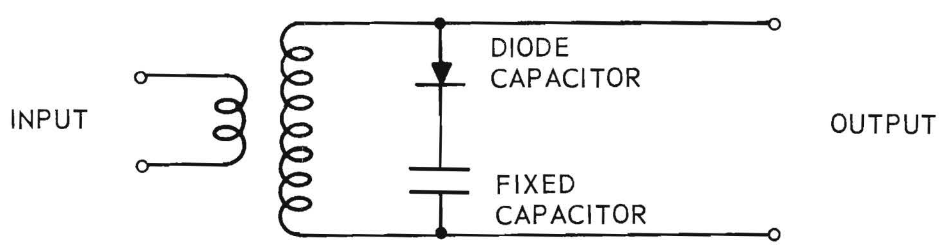



Figure 6. Resonant Circuit With Non-Linear Capacitance

of the applied signal was on the opposite side of resonance, an unstable envelope in the output was obtained. This was due to the fact that as the amplitude of the applied signal was increased, the bias voltage increased, tuning the circuit closer to the frequency of the applied signal. This resulted in an even larger bias voltage and subsequent tuning of the circuit even closer to resonance. This action is a cumulative one, so that the envelope becomes unstable, as shown in the sketch of Figure 7. This effect is undesirable in obtaining limiting action since whether or not limiting is obtained depends upon the frequency of the applied signal.

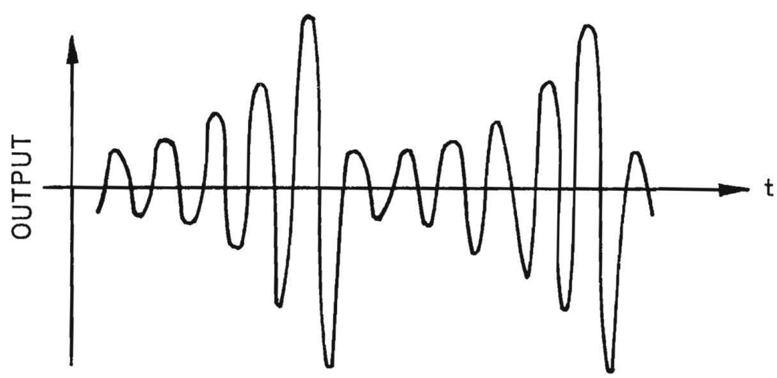



Figure 7. Envelope Instability

2.2.2 Oscillating Parametric Amplifier: The gain characteristics of an oscillating parametric amplifier offers a possible means for obtaining limiting action. Specifically, if the pump frequency is considerably higher than the signal frequency and if the desired signal output is taken at the upper sideband frequency while the output at the lower sideband frequency is terminated in a resistor, then the input-output characteristic for the desired signal will be linear, provided the pump and signal levels are sufficiently low. However, if sufficient signal or pump power is supplied to cause oscillation at the lower sideband frequency, the signal or pump energy will be used to maintain the oscillation, and the gain for the desired signal will be reduced. Thus, if the pump level is set just below the value required to cause oscillation, a very small input signal is sufficient to cause the start of oscillation and a limiting characteristic for the input signal is obtained. A simple form in which this limiting action of the parametric amplifier may be obtained is by operating the amplifier in the degenerate mode. In this mode the pump frequency is set equal to twice the tuned frequency of the amplifier. Since the idler frequency is equal to the difference between the pump frequency and the amplifier's tuned frequency it is apparent that, in the degenerate mode, the idler frequency and the amplifier's tuned frequency are equal. As a result, a single resonant circuit provides the necessary tuning for both the signal and idler frequencies. In this device the limiting action is obtained by using the signal to be limited as the pump source for the amplifier. This pump forces the amplifier into oscillation at one-half the pump frequency causing the input signal energy to be dissipated in the maintenance of the oscillation at the sub-harmonic frequency. Figure 8 illustrates one arrangement by which limiting action may be obtained. In this arrangement the output is at the same frequency as the input by means of the resonant circuit. At very low signal levels no oscillation takes place so that an essentially linear relationship exists between the input and output circuits. However, when the input signal level reaches a sufficiently large value to cause oscillation to commence at one-half the input frequency, the

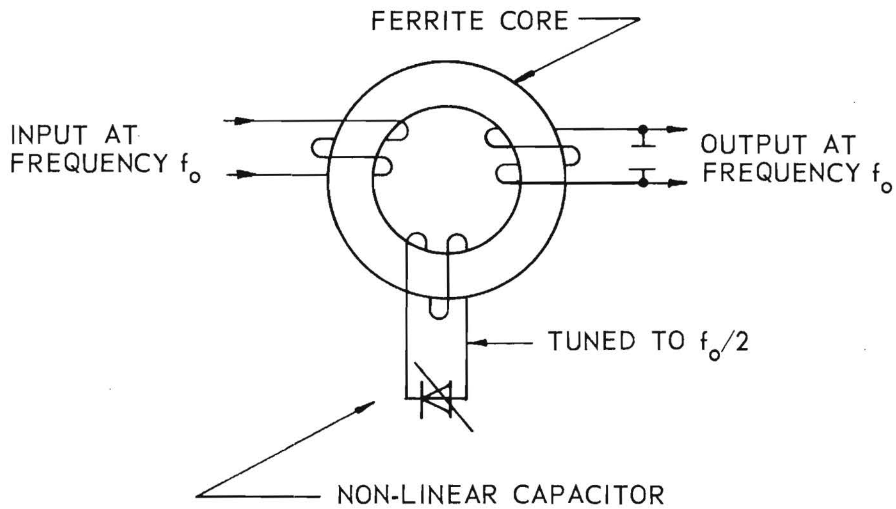



Figure 8. Limiter

input energy at levels above this value is dissipated at one-half the frequency of the input and does not appear in the output circuit. The following analysis illustrates the existence of a critical level for the start of sub-harmonic oscillation. The particular circuit to which the analysis applies is given in Figure 9. The signal level at the start of oscillation at one-half the input frequency is small so that only the linear term in the variation of capacitance with voltage is used. Since the effects of higher order terms become smaller and smaller as the signal level tends toward zero, the variation in capacitance is given by:

$$C(V_c) = C_o + C_1 V_c \ . \tag{1}$$

From the definition of capacitance,

$$C(V_c) = \frac{Q_c}{V_c} \ , \tag{2}$$

or

$$Q_c = V_c \ C(V_c) \ .$$

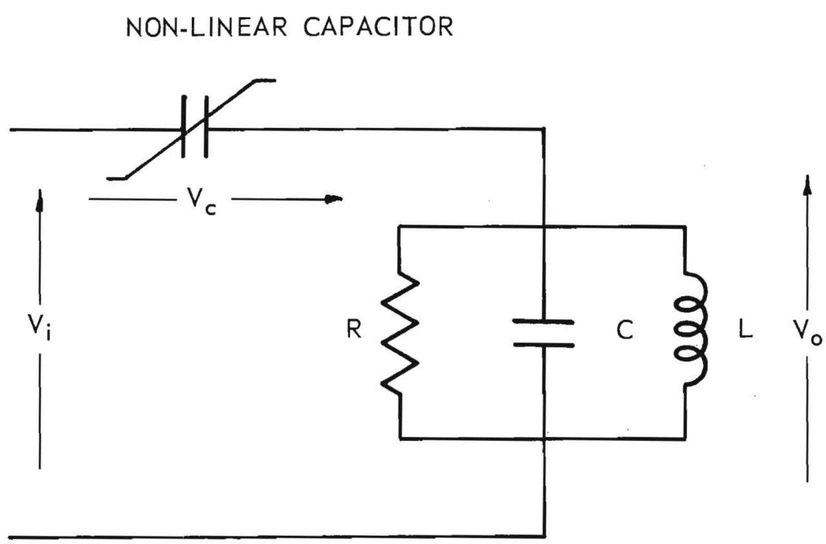



Figure 9. Nonlinear Circuit

But

$$i_c = \frac{dQ_c}{dt} = \frac{d}{dt}\left(C_o V_c + C_1 V_c^{\,2}\right) = \left(C_o + 2C_1 V_c\right) \frac{dV_c}{dt} \ . \tag{3}$$

It is assumed that the impedance of the tuned circuit is equal to R at the frequency $\omega = \omega_1/2$, and is equal to zero at all other frequencies. Since the case of interest is that concerning the start of oscillation, the components of voltage at frequency $\omega = \omega_1/2$, appearing across the non-linear capacitor will be small, so that the capacitance variation is controlled entirely by the amplitude of the input pumping signal. If the input signal in Figure 9 is,

$$V_i = A \operatorname{Sin} \omega_1 t \ , \tag{4}$$

the variational term of the capacitance is

$$\Delta C = AC_1 \operatorname{Sin} \omega_1 t \ . \tag{5}$$

Under these conditions the formula for the gain of the degenerate parametric amplifier given by Blackwell[1] is applicable. The relation is found on page 74 and is given in equation (6).

$$\text{gain} = 4\ G_g G_L \Big/ \left[ G_g + G_L - \frac{\omega_1 \omega_2 (\gamma\ C_o)^2}{G_g + G_L} \right]^2 \tag{6}$$

This gain becomes infinite for:

$$(G_g + G_L) - \frac{\omega_1 \omega_2 (\gamma\ C_o)^2}{G_g + G_L} = 0 \tag{7}$$

or:

$$(G_g + G_L)^2 - \omega_1 \omega_2 (\gamma\ C_o)^2 = 0 \tag{8}$$

or:

$$(G_g+G_L)^2 = \omega_1\omega_2(\gamma\ C_o)^2 \ . \tag{9}$$

In terms of the circuit of Figure 9,

$$\left.\begin{aligned} G_g+G_L &= 1/R \\ \omega_1 &= \omega_2 \\ C_o &= C_1 \\ \gamma &= \frac{A}{2} \end{aligned}\right\} \tag{10}$$

so that for this circuit the condition for oscillation reduces to

$$1/R^2 = \omega_1{}^2A^2C_1{}^2/4 \ ; \tag{11}$$

solving for A gives:

$$A = 2/\omega_1C_1R \ . \tag{12}$$

This result indicates that the input signal level, A, must exceed the value $2/\omega_1C_1R$ for oscillation and hence limiting to begin.

The appearance of the quantity $\omega_1$ in the denominator of equation (12) indicates that the amplitude required for limiting to begin decreases at higher frequencies. As a result, it would be expected that such limiters would find their greatest usefulness at higher frequencies.

To test these results, the circuit of Figure 9 was constructed in the laboratory and the threshold of oscillation was determined to be approximately 0.3 volts. The following values for the components were measured:

$$C_1 = 2.5 \times 10^{-12}$$

$$R = 10^3$$

$$\omega_1 = 2\pi \times 2.5 \times 10^8.$$

Substituting these values in equation (12) gives a value of 0.5 volts which is in reasonable agreement with the measured value of 0.3 volts. However, this value is much too large to provide an appreciable protection when used at the input terminals of a receiver. Unfortunately, the level of interfering signals commonly encountered is considerably below 0.3 volts. In an attempt to lower the value of this limiting level, the limiter of Figure 10 was constructed. In this arrangement some increase in the signal voltage applied to the nonlinear capacitance is obtained through the impedance transformer of the 250 mc resonant circuit, however no appreciable lowering of the limiting threshold was obtained although the desired limiting action was observed. As a result, it is felt that although these parametric limiting devices provide the desired receiver protection at very large signal levels they are not generally satisfactory for the protection of receivers against low level pulse interference.

### 2.3 Spurious Response Reduction

Many currently used measurements receivers suffer from poor rejection of spurious responses. These responses greatly complicate the task of making accurate and reliable measurements since the spurious responses of the

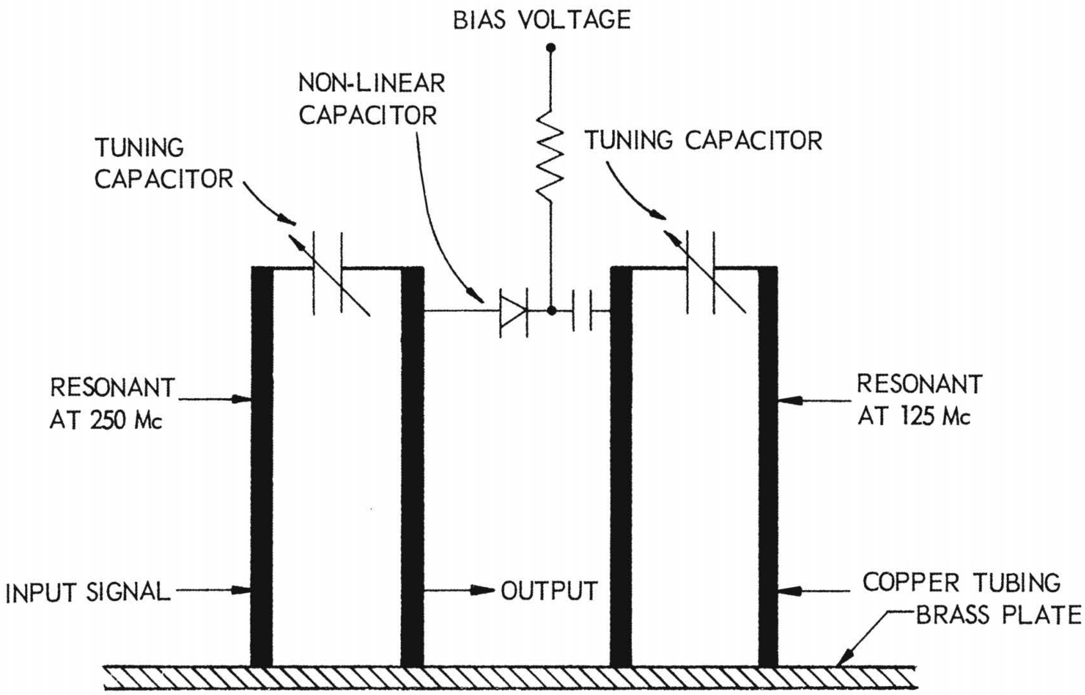



Figure 10. Parametric Limiter

measurements equipment may mask or be mistaken for spurious emissions of the equipment under test. Although techniques have been developed to permit the identification of the responses which are associated with the measurements equipment, the elimination of these responses represents a more desirable solution to this problem. In order to illustrate some techniques which are available to eliminate or reduce the magnitude of these responses, an adaptor was constructed for use in front of existing equipment to provide an improvement in the spurious response rejection of this equipment.

This adaptor contains a collection of techniques which are effective in improving the spurious response rejection of receiving equipments. These techniques are (1) a cavity preselector, (2) a linear RF amplifier, (3) a low spurious response balanced mixer, and (4) a local oscillator system in which the harmonic output has been drastically reduced. These separate devices have been arranged in such a manner that external connection may be made to each of them by means of front panel jacks. As a result, any desired combination of these devices can be used to illustrate the improvement in spurious response rejection obtained in a given situation.

2.3.1 Cavity Preselector: A portion of the spurious response rejection of the receiver adaptor is provided by a tunable cavity preselector covering the 200 to 400 mc tuning range of the equipment. In the interest of conserving space, a considerable amount of capacitance loading is used at the lower frequency end of the tuning range of the cavity. Sufficient capacitance loading is provided to permit the resonant frequency to be adjusted below 200 mc with a physical length of only 6 inches. Figure 11 illustrates the conical shaped top loading arrangement which has been placed at the top of the center

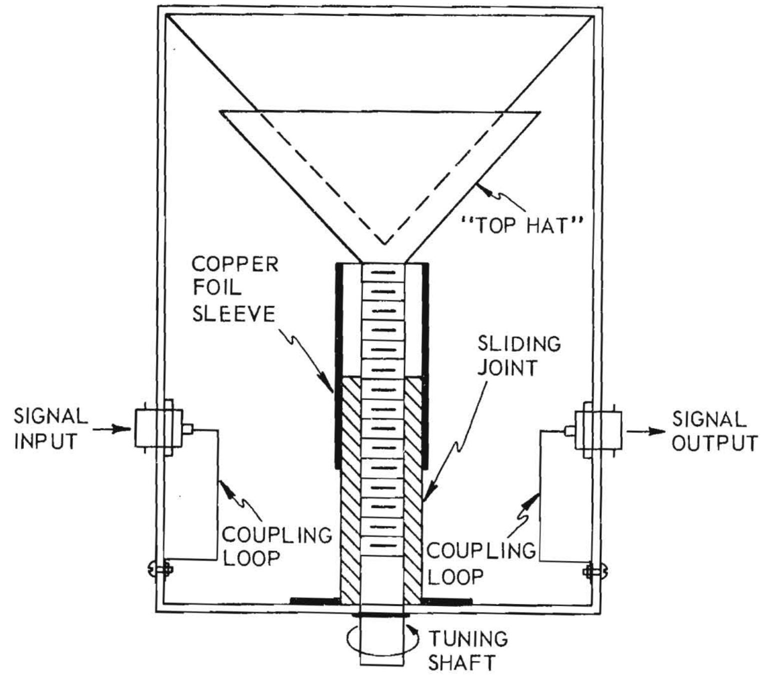



Figure 11. Cavity Preselector

conductor. This particular shape was chosen to provide a reasonably constant tuning rate over the entire tuning range of the filter. If the top loading had been provided in the form of a simple disc-shaped top hat then the capacitance would change very rapidly at the lower end of the tuning range and the tuning rate in this region would be much too fast. The use of the cone shaped top hat permits the change in capacitance to be more gradual with a consequent reduction in the low frequency tuning rate. Figure 12 illustrates the tuning curve obtained for two values of the cone angle. The cone angle labeled 60 degrees was chosen for use in the final model of the cavity preselector and provides an almost constant tuning rate across the entire tuning range. Input and output coupling to the cavity is by means of a pair of inductive coupling loops, the sizes of these loops being adjusted to provide the minimum insertion loss con-

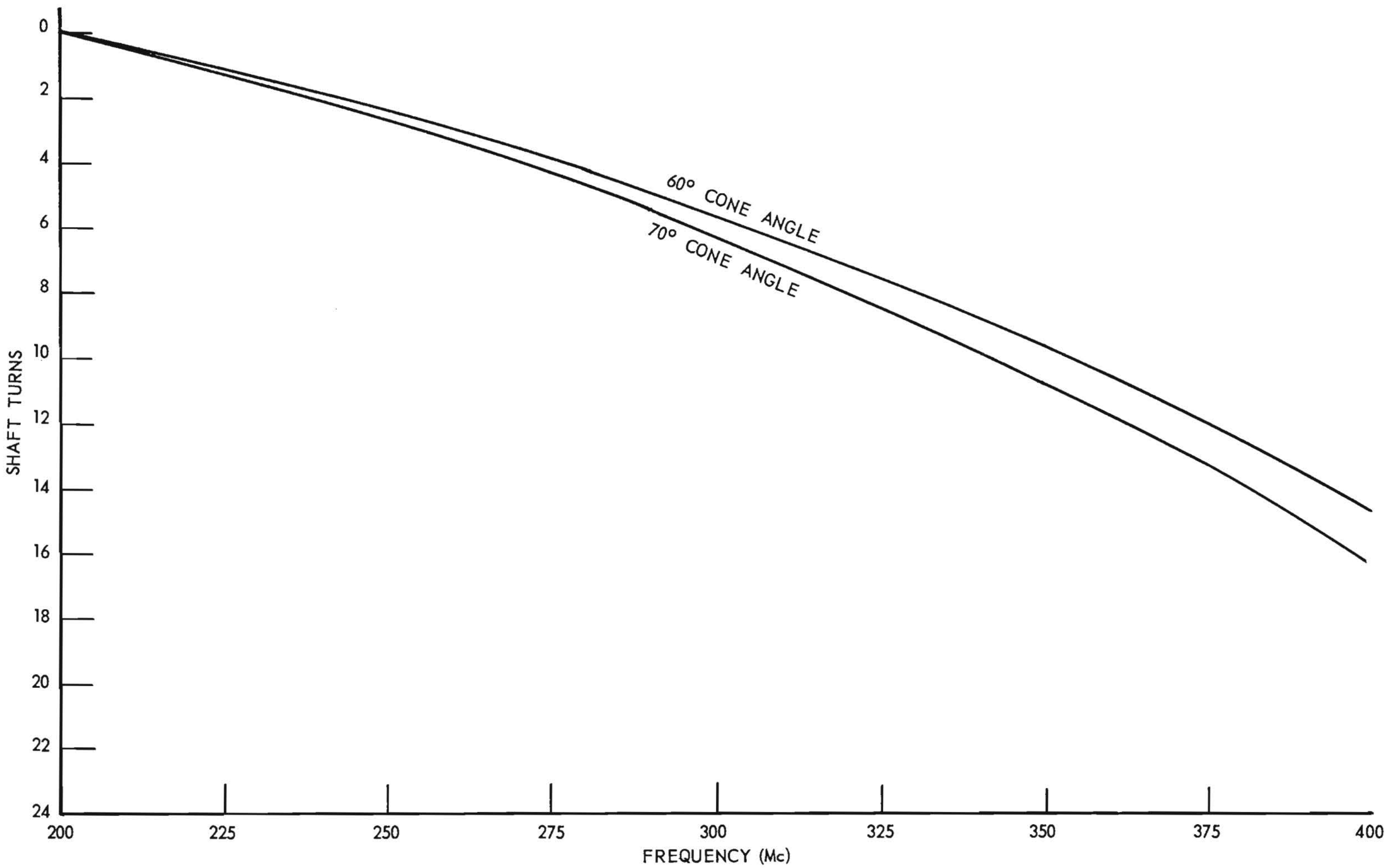



Figure 12. Effect of Cone Angle on Tuning Curve

sistent with obtaining the desired selectivity. The curve of Figure 13 shows the insertion loss of the cavity resonator at its tuned frequency as the tuned frequency is varied from 200 to 400 mc. Notice that this insertion loss is 1 db or less over the entire range. A typical selectivity curve for this cavity is given in Figure 14. Notice that the skirt selectivity is somewhat poorer on the low side of the resonator than on the high side of the resonator. This asymmetry of the selectivity curve results from the large amount of capacitive loading used in resonating the cavity. The selectivity curve of Figure 15 illustrates the results obtained when two of these cavities are cascaded and tuned to the same frequency. Since the tuning is quite sharp, it is necessary that the two cavities be machined quite carefully so as to make possible the accurate tracking of one cavity with the other. Such an arrangement provides a high selectivity preselector for the range 200 to 400 mc in a very small physical space.

Another advantage obtained by the use of a large amount of capacitance top loading of the resonator cavity is the reduction of the amplitudes of the spurious responses normally associated with such structures. In those cavity resonators which do not utilize capacitance loading, pass bands are encountered in the neighborhood of those frequencies at which the length of the center conductor of the coaxial cavity is equal to an odd multiple of 1/4 wave length. However, when capacitance loading is employed, a resonance is encountered at those frequencies for which the equivalent inductive reactance of the coaxial cavity, viewed as a short circuited transmission line, is equal to the capacitive reactance of the loading capacitor. Figure 16 illustrates graphically the variation of the resonator reactance and the

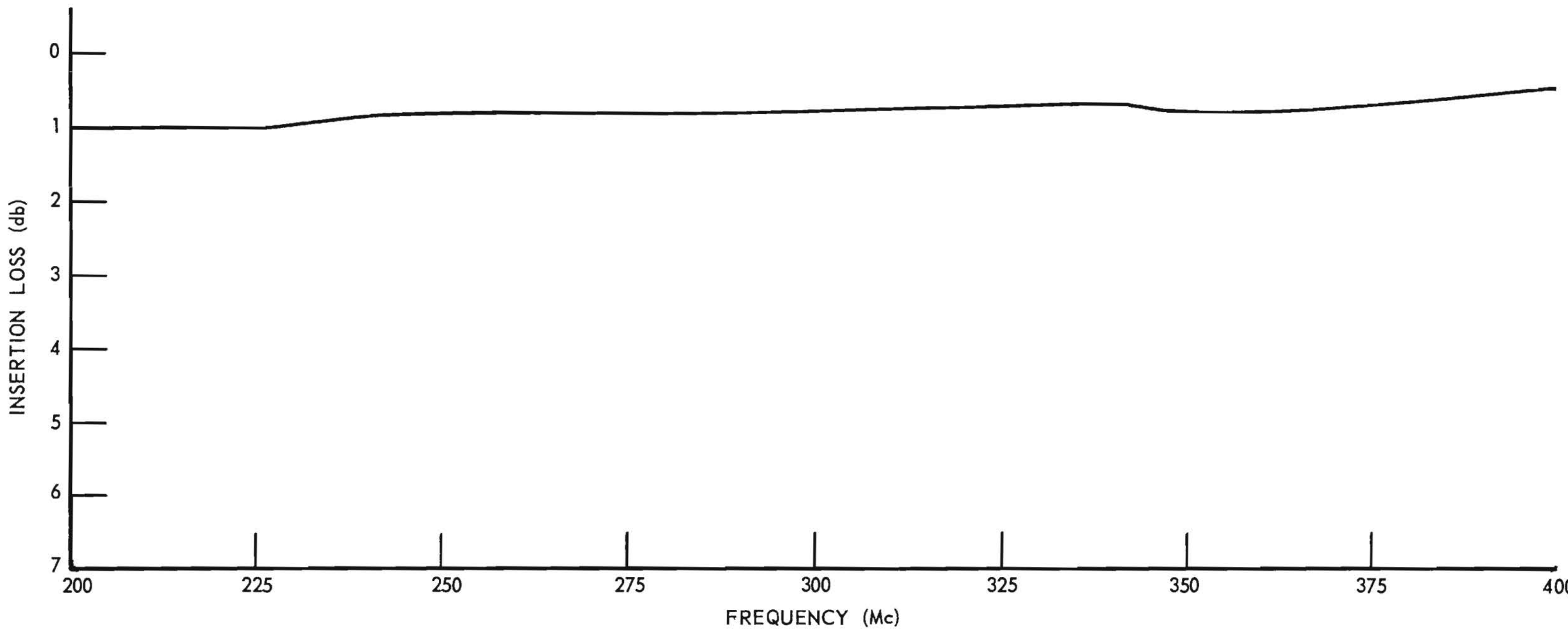



Figure 13. Cavity Resonator Insertion Loss

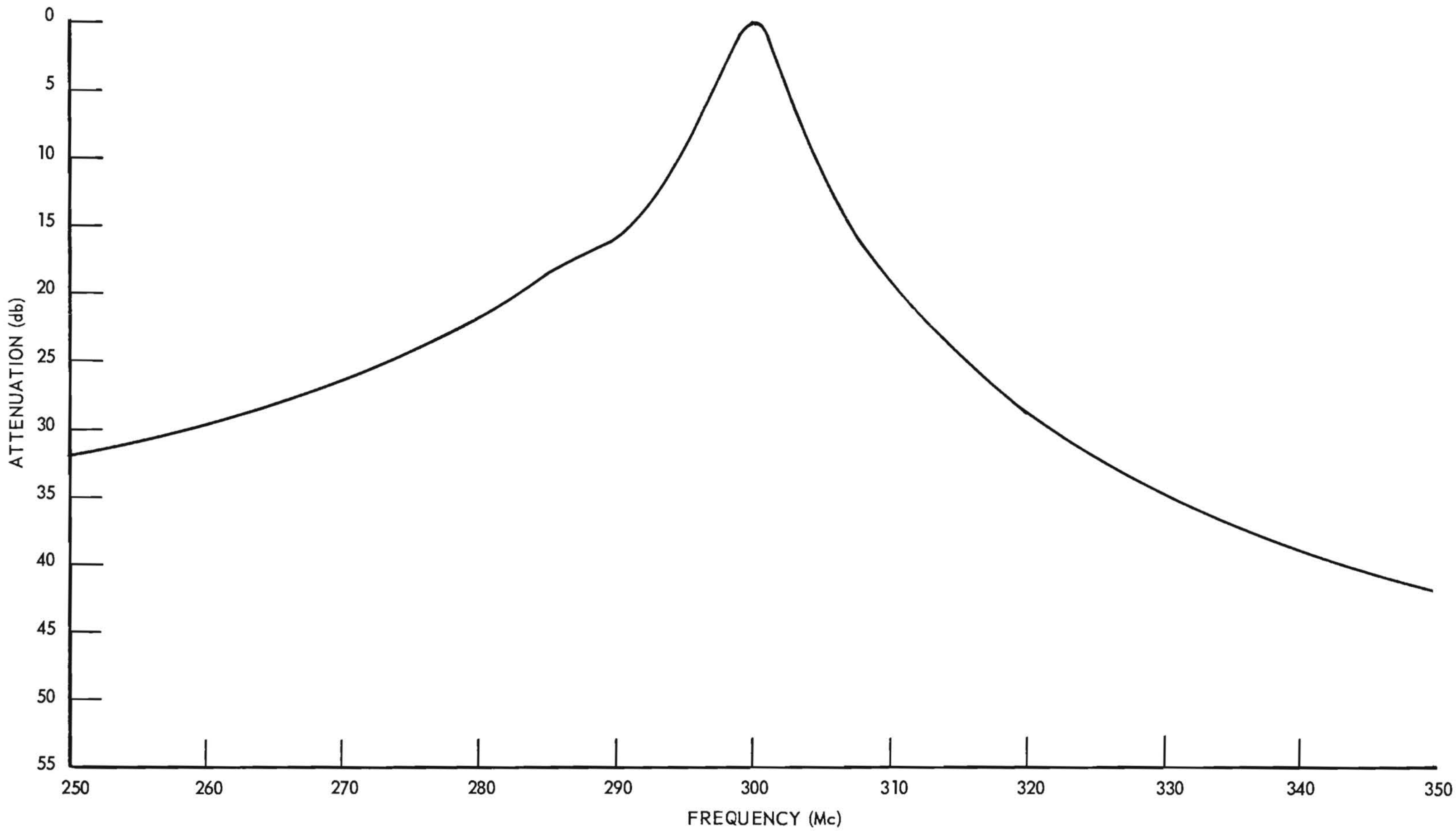



Figure 14. Single Resonator Selectivity Curve

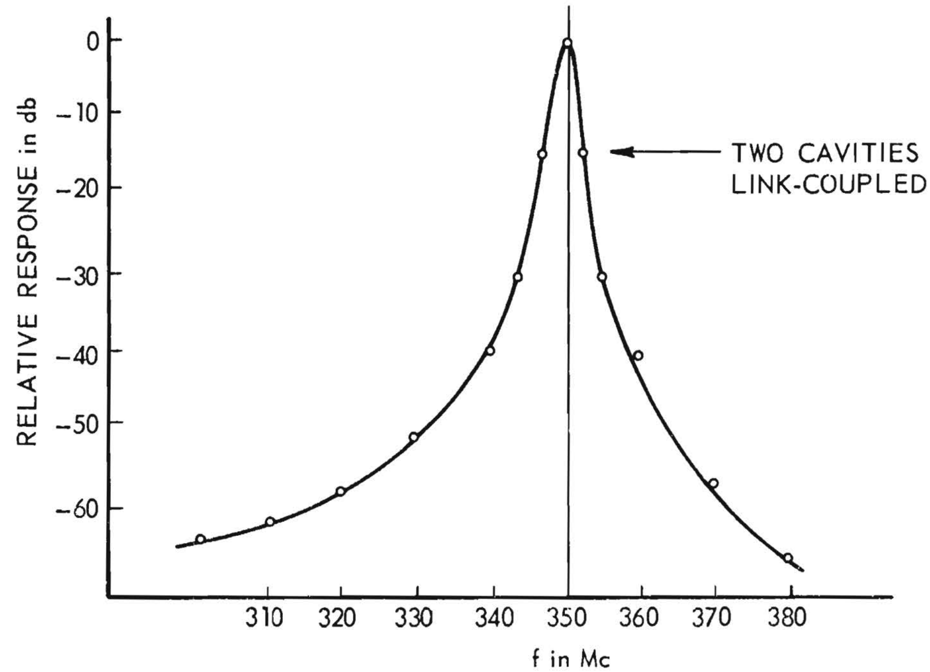



Figure 15. Two Cavity Selectivity Curve

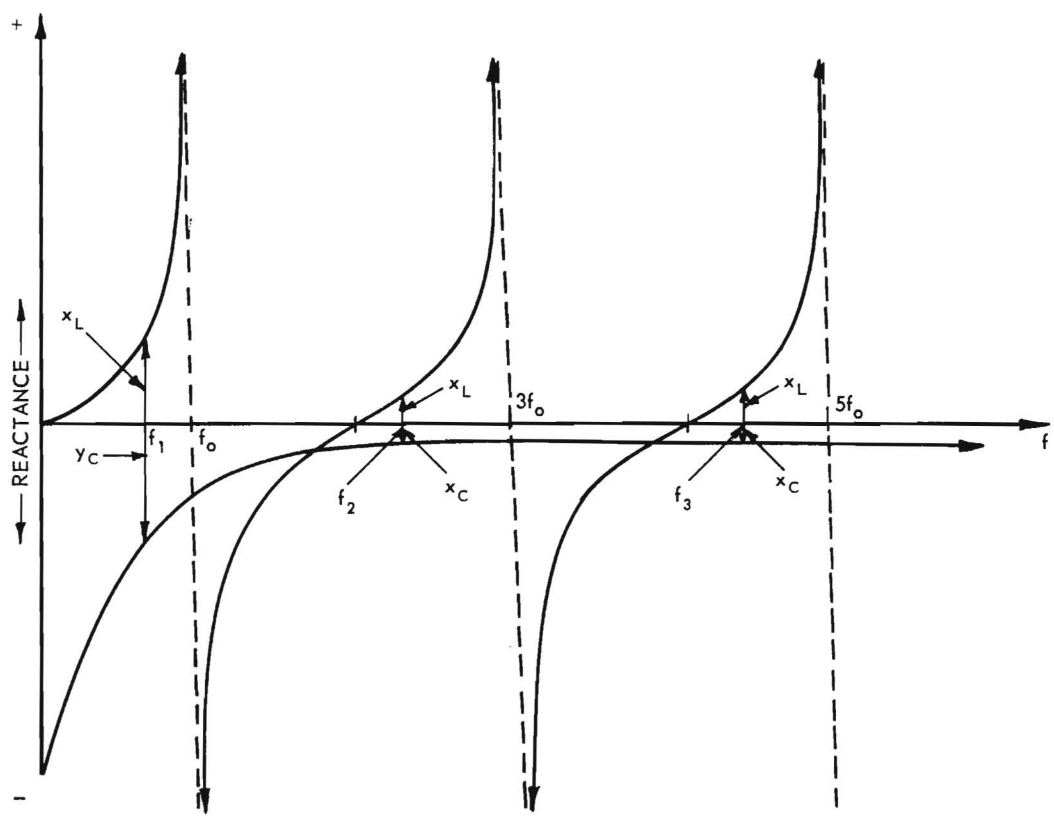



Figure 16. Cavity Reactance Functions

reactance of loading capacitance as a function of frequency. The frequency locations at which the two reactance functions are equal are indicated on the figure. Two characteristics of these frequencies should be noted. First is the fact that these frequencies are not harmonically related and, secondly, the magnitude of the inductive reactance and hence the "Q" associated with a particular resonance falls off rapidly as the frequency is increased. This low "Q" causes the amplitudes of these responses at the higher frequencies to be considerably reduced over those encountered in resonators not employing capacitance loading. The curve of Figure 17 illustrates the response of a single cavity tuned to 300 mc as the frequency of the driving source is varied for 100 mc to 4 KMC. The lack of the large response usually occurring near the 3rd harmonic of the tuned frequency should be noted. If this cavity filter is cascaded with a low pass filter whose cutoff lies in the neighborhood of 1000 mc, the spurious response problem can be almost completely avoided since the low pass filter will suppress the response of the cavity above the cutoff frequency of the low pass filter.

2.3.2 RF Preamplifier: An RF Preamplifier is included in the receiver adaptor to provide additional sensitivity and to establish a good skirt selectivity to avoid the spurious responses of the cavity preselector which occur at out-of-band frequencies. The physical construction of this preamplifier is illustrated in the photo of Figure 18 and a schematic is shown in Figure 19. This preamplifier consists of two 6J4 grounded grid amplifier stages which provide a gain of about 20 db over the 200 to 400 mc band. Two tuned circuits are incorporated to provide the previously mentioned spurious response selectivity and to permit the necessary inter-stage impedance level adjustments to be made.

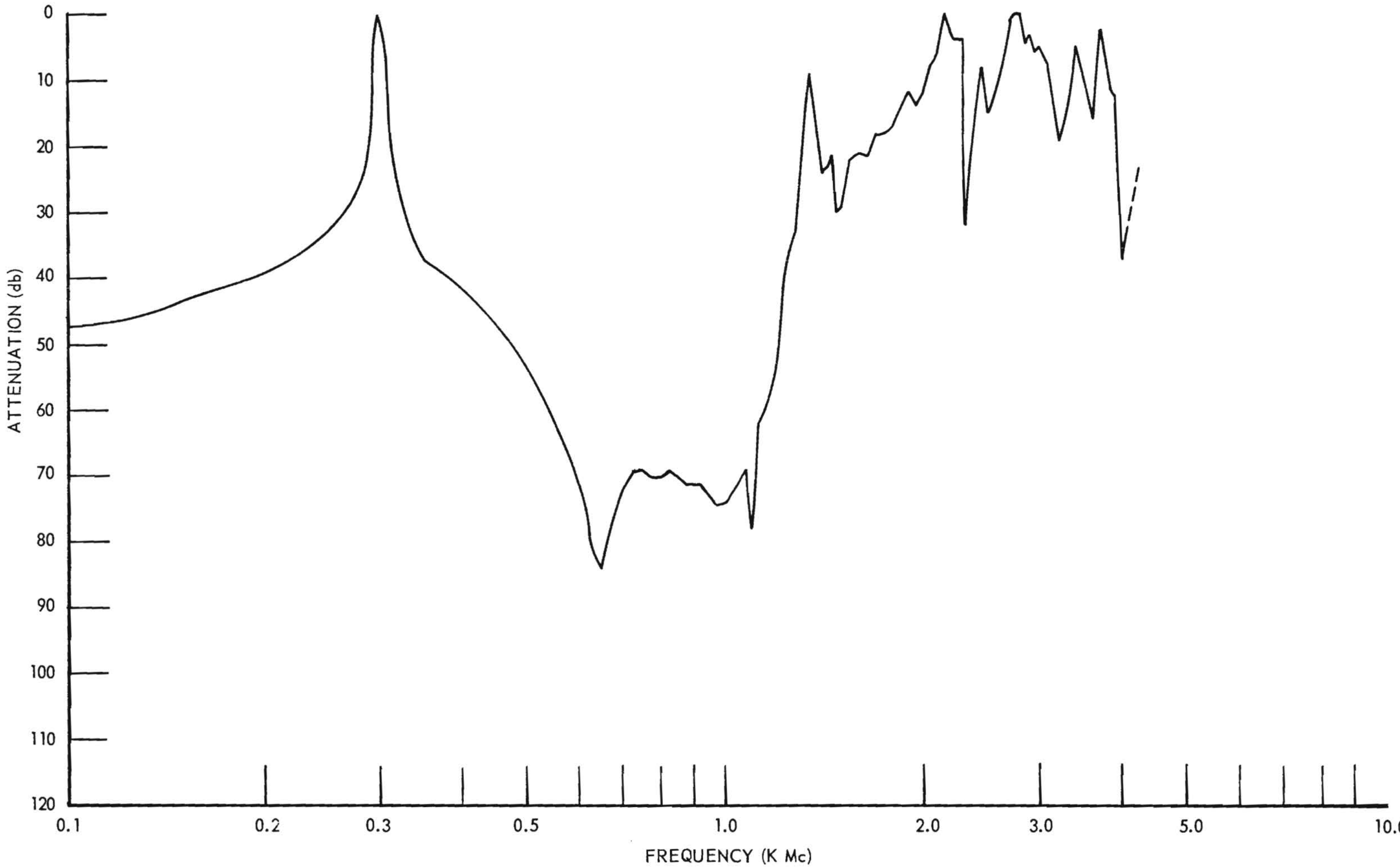



Figure 17. Cavity Spurious Responses

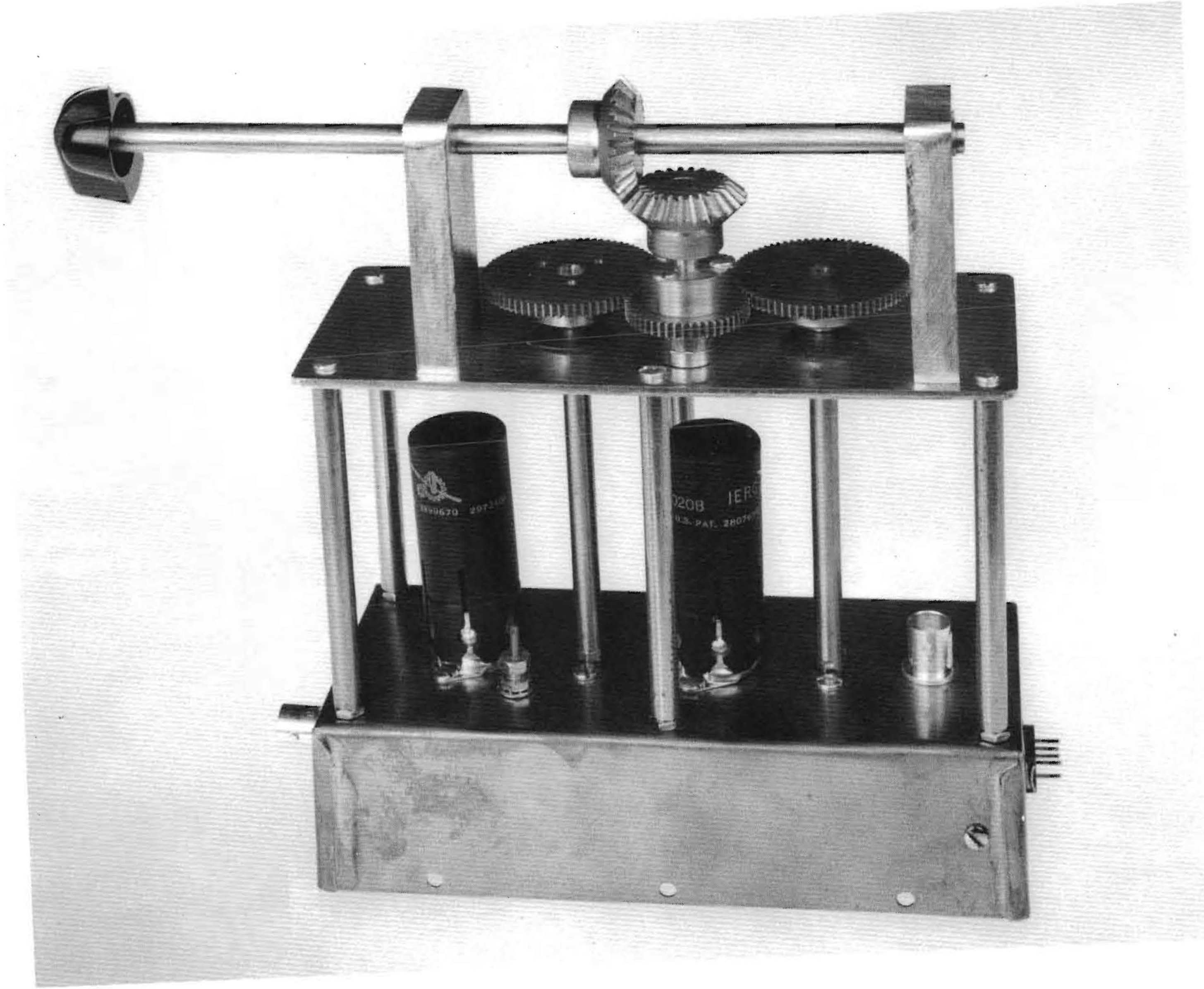


Figure 18. Outside View of RF Preamplifier

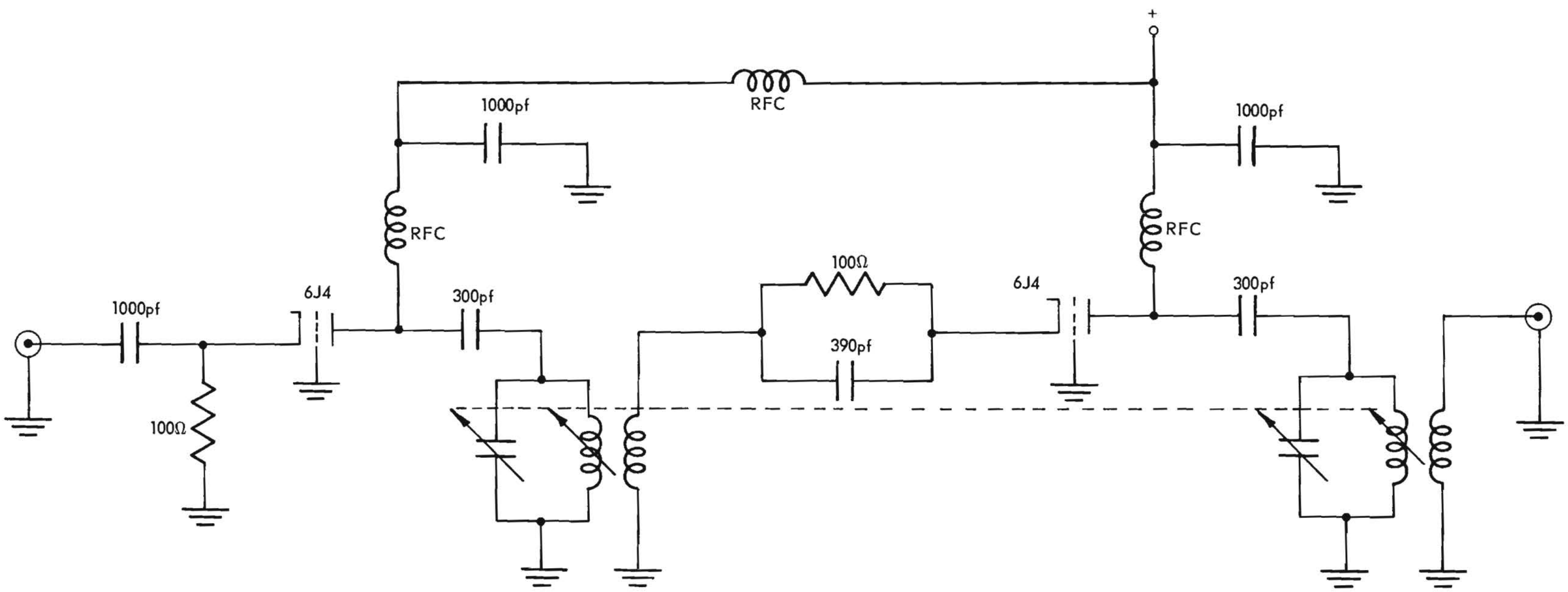



Figure 19. RF Preamplifier Schematic

A mechanical tracking arrangement consisting of a gear drive from a single control shaft has been constructed to permit single knob control of both tuned circuits. The tracking error associated with this tuning mechanism reduces the available gain to approximately 17 db, but this loss is not considered significant since the net gain is more than adequate to establish a satisfactory noise figure for the entire preamplifier and mixer combination. Addition of this selectivity to that already obtained in the cavity preselector produces approximately 110 db rejection on the skirts of the selectivity curve.

Some difficulty was encountered in obtaining a 200 to 400 mc tuning range. This arises primarily from the fact that the minimum capacitance values associated with the circuitry limit the available capacity ratio to less than 4 to 1. As a result, a 2 to 1 tuning range could not be obtained by variation of the capacitor alone. To overcome this difficulty, a modified capacitor was constructed with a slider attached to the rotor plates of the variable capacitor. This slider shorts out a portion of the inductance as the tuning shaft is rotated so that a large variation in inductance as well as capacity is obtained. With this arrangement, the desired 2 to 1 tuning range was easily achieved. The mechanical details of this arrangement can be seen in the photograph of Figure 20.

Figure 21 illustrates the gain of the amplifier as a function of frequency while Figure 22 illustrates the amount of overall selectivity obtained by the use of two tuned circuits.

2.3.3 Mixers: Since the mixer is the major offender in generation of spurious responses, a considerable amount of effort was devoted to the

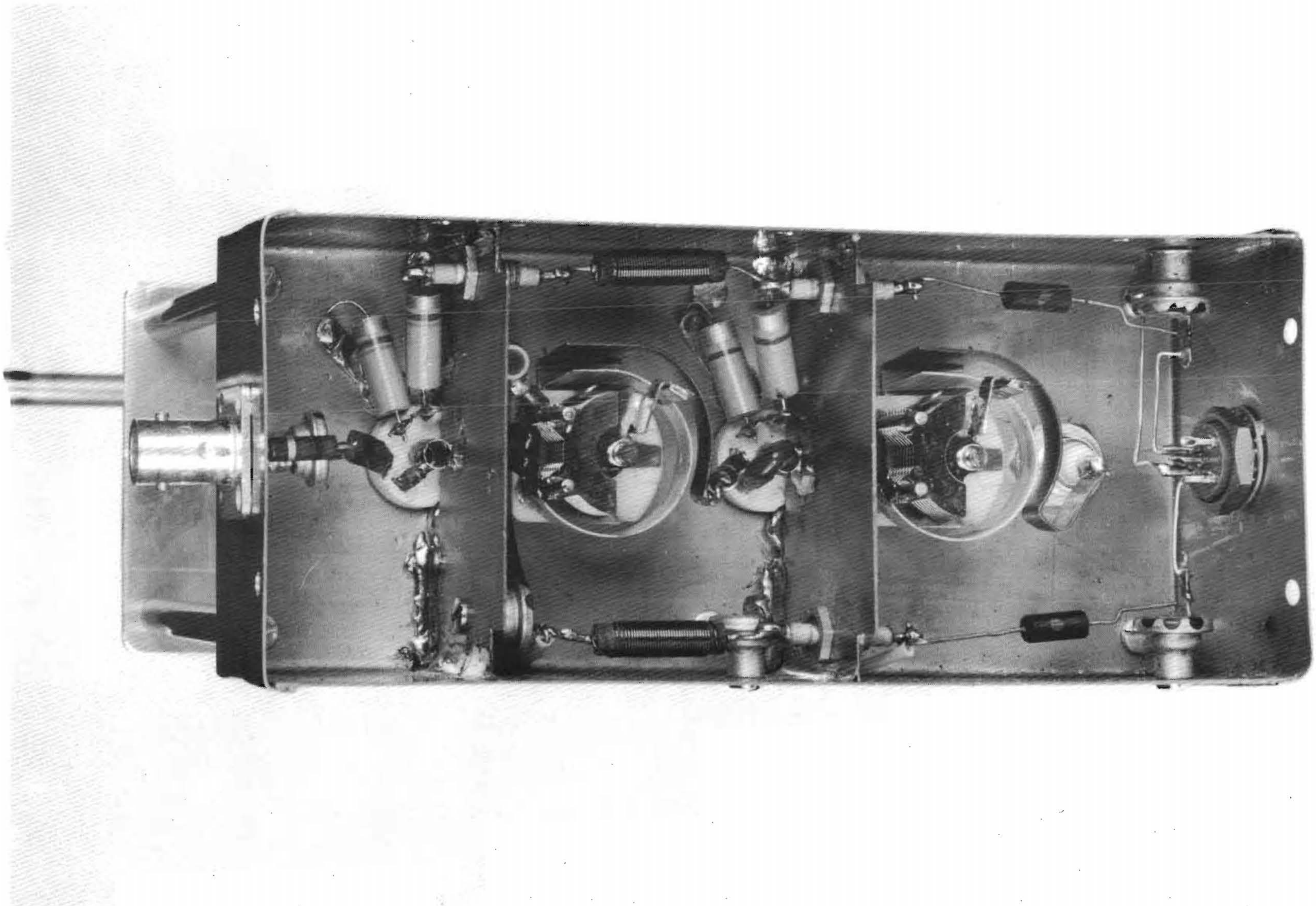


Figure 20. Internal View of RF Preamplifier

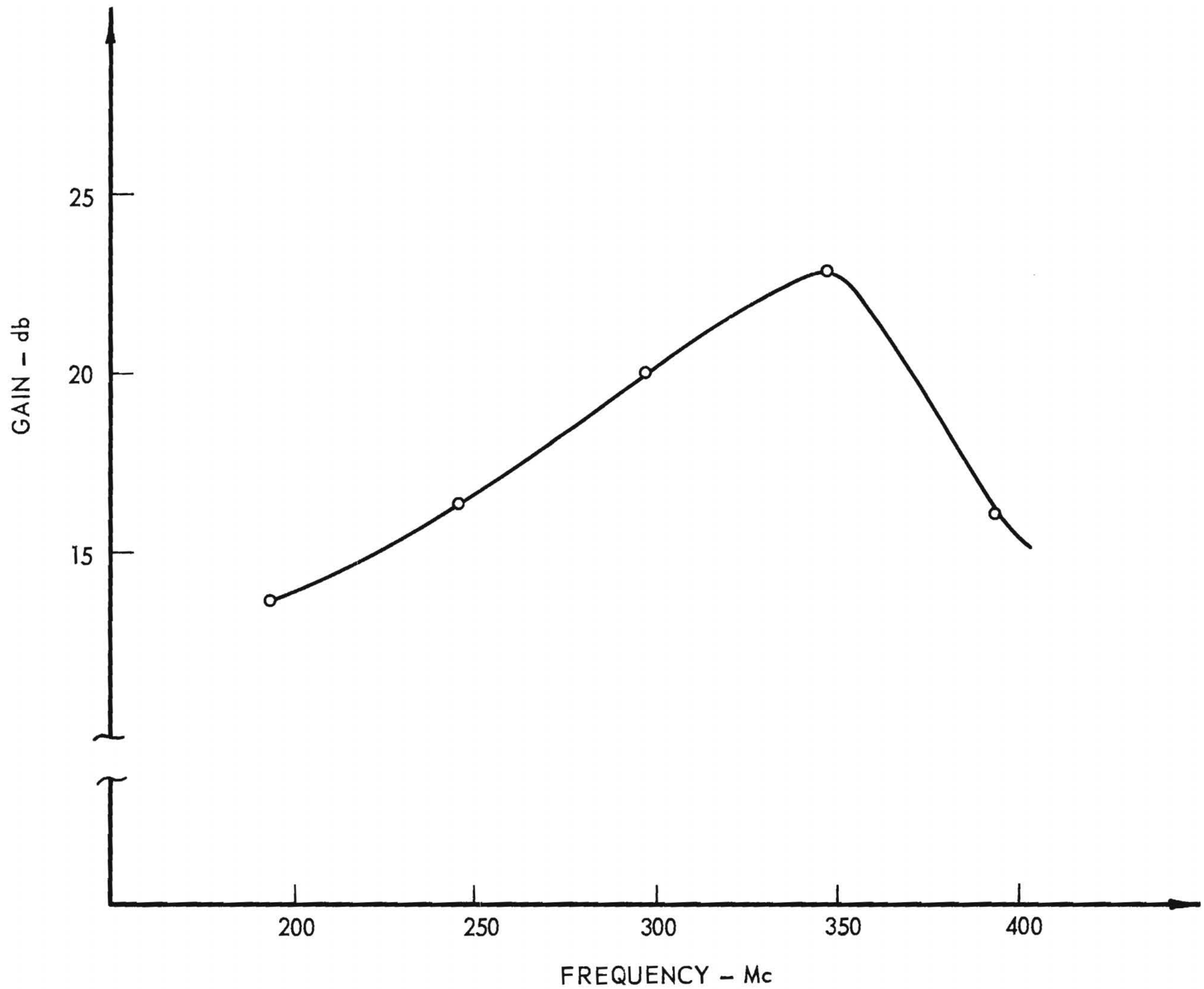



Figure 21. Preamplifier Gain versus Frequency

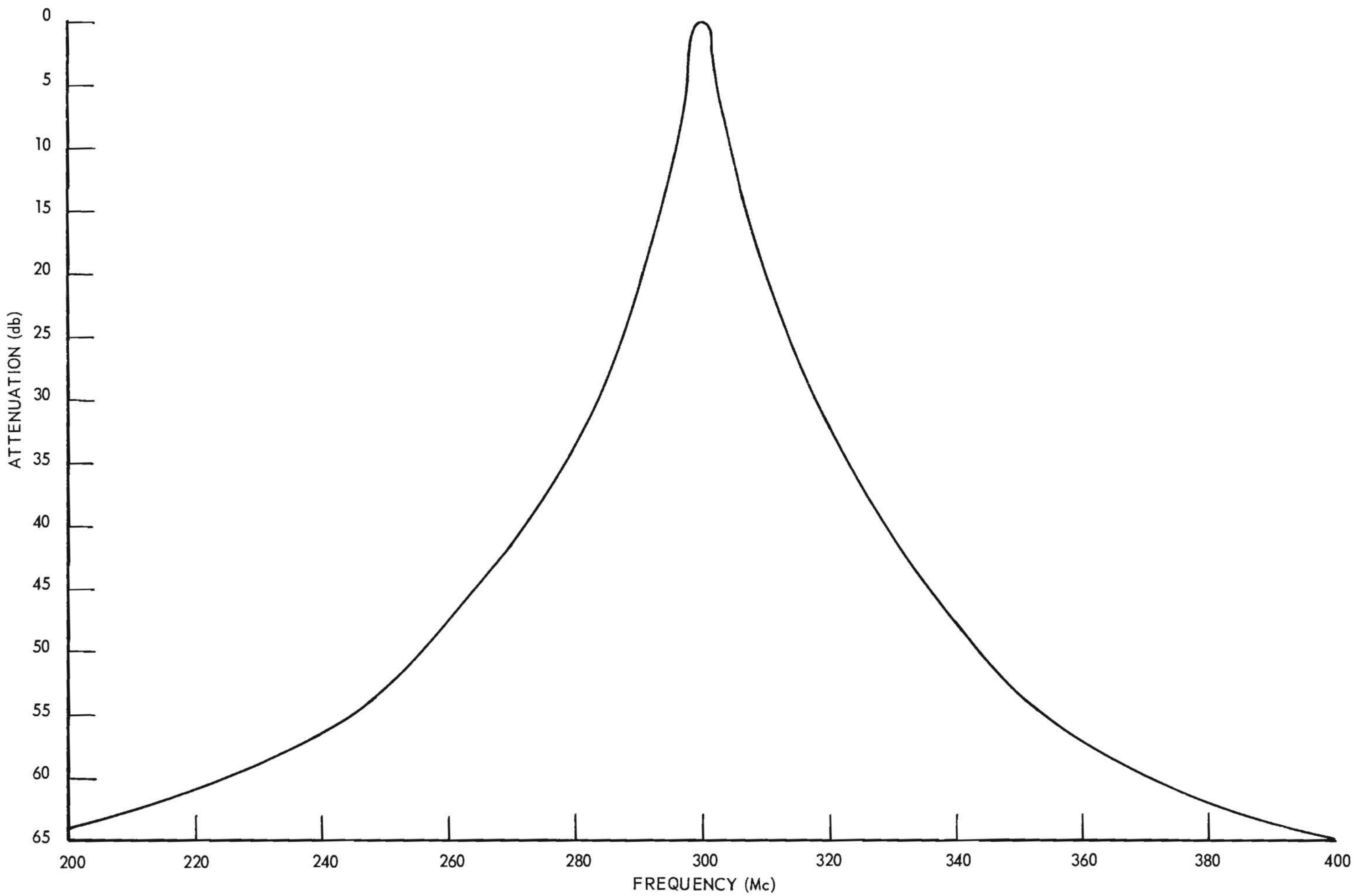



Figure 22. Preamplifier Selectivity Curve

investigation of methods for the optimization of mixers with respect to spurious response reduction.

Although the frequency locations of the spurious responses are the same for all mixers, the magnitudes of these responses may vary widely from one mixer to the next; the variation being dependent upon the detailed nonlinear characteristics of the mixers. It is by careful adjustment of these nonlinear characteristics that an improvement in spurious response rejection of a particular mixer can be obtained.

2.3.3.1 Operation at Reduced Signal Levels: One technique by which spurious responses may be reduced is to maintain the signal and local oscillator drive level at the mixer terminals at the minimum value consistent with usable operation. The success of this method depends upon the fact that when a power series expansion of the nonlinear mixer characteristic is made, it is found that the spurious responses arise from the higher powers in the expansion. Consequently, a reduction in the level of the signal which gives some particular decrease in the desired signal level will give a considerably greater decrease in the level of the undesired response since this undesired response is proportional to a higher power of the signal amplitude than that of the desired response. For example, a reduction in the input signal by a factor of two will reduce a spurious response arising from the fifth harmonic of the input signal by a factor of $2^5$ or 32, while reducing the desired response only by a factor of two. This represents an improvement in the ratio of desired to undesired of 16 to 1, or 24 db. In many instances, the reduction of the response to the desired signal, that is a lowering of the conversion gain of the mixer, does not constitute a serious problem and is a very cheap price to pay for the greatly improved rejection of spurious signals. In particular, whenever the signal-to-noise ratio has been adequately established prior to the mixer either by sufficient low noise RF amplification or by the fact that the level of atmospheric noise with respect to receiver noise is quite high, then any conversion loss in the mixer can be easily overcome by a corresponding increase in the gain of the IF amplifier.

It should be noted that the IF amplifier may exhibit some degree of nonlinearity and it is important to provide as much of the IF selectivity as possible before any amplification takes place. This avoids the possibility that some spurious responses may be generated in the nonlinearity of the IF amplifier.

A simple method of insuring that these nonlinear products are not formed in the IF amplifier is to provide the selectivity in the form of a crystal or mechanical filter, operating directly at the mixer output terminal. The current availability of these filters possessing superior selectivity characteristics make this solution to the problem very attractive. In reference to crystal filters, designers have been able to provide the normally required IF bandwidths at center frequencies, ranging upwards of 50 mc, thereby obtaining the high image rejection which can be realized by the use of a high IF frequency while at the same time maintaining the simplicity of the single conversion receiver.

2.3.3.2 Selection of Operating Point: An alternative to the operation of a mixer at a very low drive level with the consequent low conversion gain is to carefully select the operating point so as to minimize the undesirable nonlinear characteristics which are responsible for the existence of spurious responses, while at the same time attempting to maximize those nonlinear properties which are responsible for the conversion of the desired signal. Certainly, complete success in this regard is not possible, but a considerable improvement can be obtained over that type operation in which the operating point is not selected with respect to minimizing spurious responses. The existence of an optimum operating point is indicated by a consideration of the Taylor's series expansion of the plate current versus grid voltage characteristics of the mixer tube about an arbitrary operating point. The coefficients in this series are proportional to the derivatives of the plate current with respect to grid voltage at the operating point. The curves of these various derivatives for commonly used mixer tubes have both maxima and minima. Since these curves are obtained by repetitive differentiation of the plate current-grid voltage curves, the maxima of even order derivatives will, in general, occur in the vicinity of the minima for odd order derivatives.

The possibility of such operation may be seen from an examination of the curves shown in Figure 23. In this figure, the first six derivatives of the plate current with respect to grid bias voltage for a typical triode have been plotted against the bias voltage. Notice that selection of the bias point at a maximum of second order curvature will give a minimum of spurious response output due to the odd order curvature. Since the desired response is proportional to the second order curvature in the mixer, a maximum desired response occurs at this bias point. That is to say that the second order curvature is responsible for forming the product of the local oscillator signal

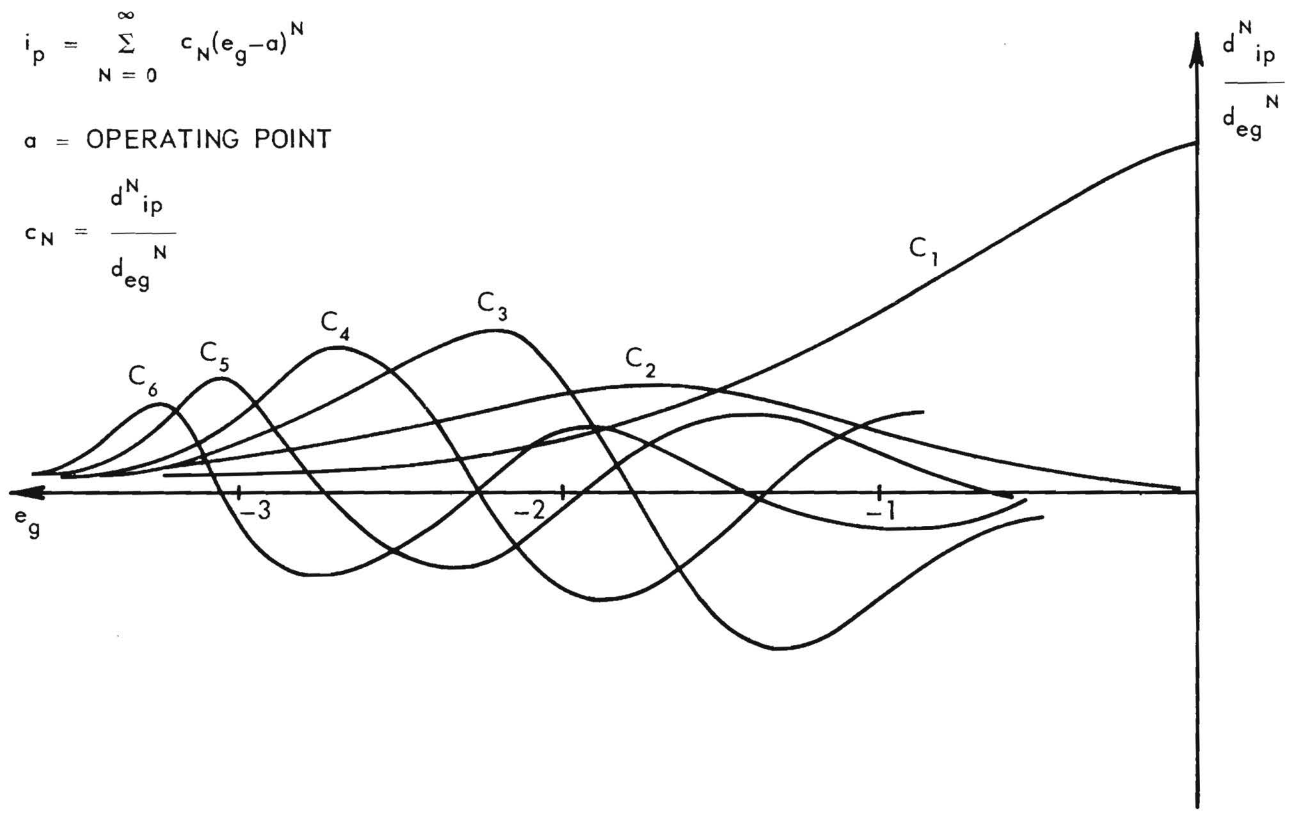



Figure 23. Derivatives of Plate Current versus Grid Voltage

and the input signal to produce the difference frequency as an IF output.

Some typical spurious response variations obtained by changing the bias point of a mixer are shown in Figure 24. One of these curves represents an even order response while the other represents an odd order response. Notice that the minima and maxima of these responses interlace each other as the grid bias voltage is varied. From the curves it can be seen that it is important, for optimum rejection of spurious responses, that the bias voltage on the mixer should be maintained at its optimum value and no automatic gain control voltage should be applied to this stage. The application of AGC voltage would cause the operating point, and, hence, the spurious response rejection, to vary with the amplitude of the AGC voltage. Adequate gain control may be obtained by control of other stages in the receiver so that it does not present any particular hardships to the designer in omitting the gain control from this stage. In this regard, it might be well to mention that the curves of Figure 23 also indicate that there is an optimum operating point for tubes used as straight amplifiers so that it would seem to be unwise to apply AGC voltage to the RF stages as well as the mixers, in order that they may be operated at an optimum with respect to the suppression of interference. In this case, it is necessary to provide some automatic gain control device directly at the antenna terminals of the receiver in order that the RF stages may not be overloaded with strong signals. Additional AGC action can also be obtained by applying AGC voltage to the IF amplifier so that the full burden of AGC actually is not placed upon whatever gain control element is used in the receiver input. One such device for accomplishing this function is described in the litera-

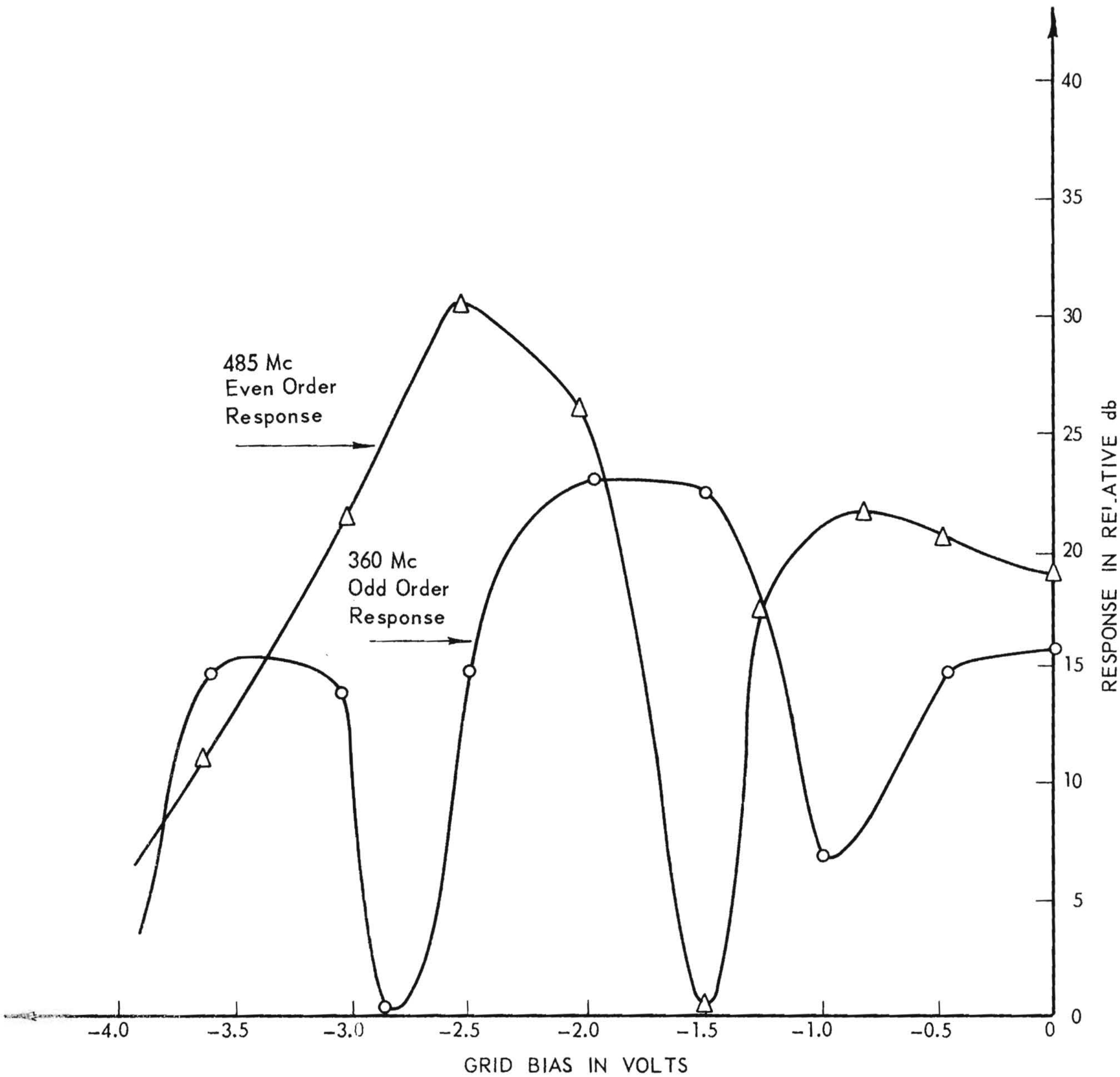



Figure 24. Spurious Response Level versus Grid Bias

ture.[2] This device is a current controlled attenuator in which the saturation of a ferrite core transformer is used to provide a signal attenuation level which is a function of the control current. The minimum loss is claimed to be quite low, in the order of 1 to 2 db with attenuation at saturation of 30 to 40 db being obtained.

2.3.3.3 Balanced Mixers: Another means by which a reduction in the level of spurious responses can be obtained is by the use of a balanced mixer. A typical balanced arrangement is shown in Figure 25. This mixer is arranged to permit cancellation of the responses due to mixing of signals with even order harmonics of the local oscillator. If the local oscillator signals to the two separate mixers are fed $180^{\circ}$ out of phase and if the IF outputs of the two mixers are subtracted to obtain the total IF output, those IF components which result from mixing with the fundamental and odd harmonics of the local oscillator will appear at their respective mixer output terminals with a $180^{\circ}$ phase difference, so that subtraction of these two signals results in the addition of the two IF components. However, those IF outputs which are due to mixing of the input signal with even harmonics of the local oscillator appear at the outputs of their respective mixers in phase, so that subtraction of these two outputs results in no net IF contribution due to these signals.

A comparison of the performance of a balanced and an unbalanced mixer is given in Figure 26. The results are shown only for even order local oscillator harmonics since no improvement is obtained for odd order responses. Notice that a general improvement in the rejection of even order spurious responses of approximately 20 db can be obtained by proper balanced mixing. Careful adjustment of the mixer circuitry makes possible larger improvements in

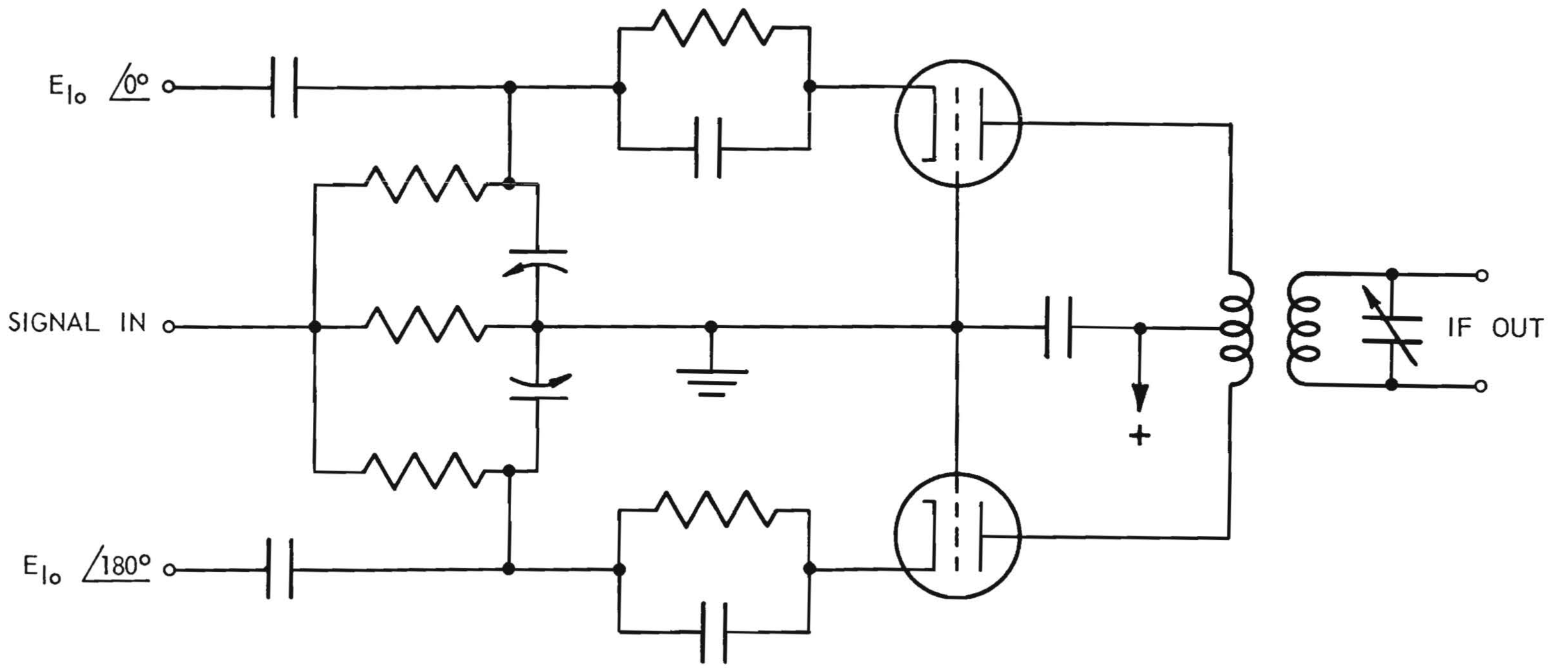



Figure 25. Balanced Mixer

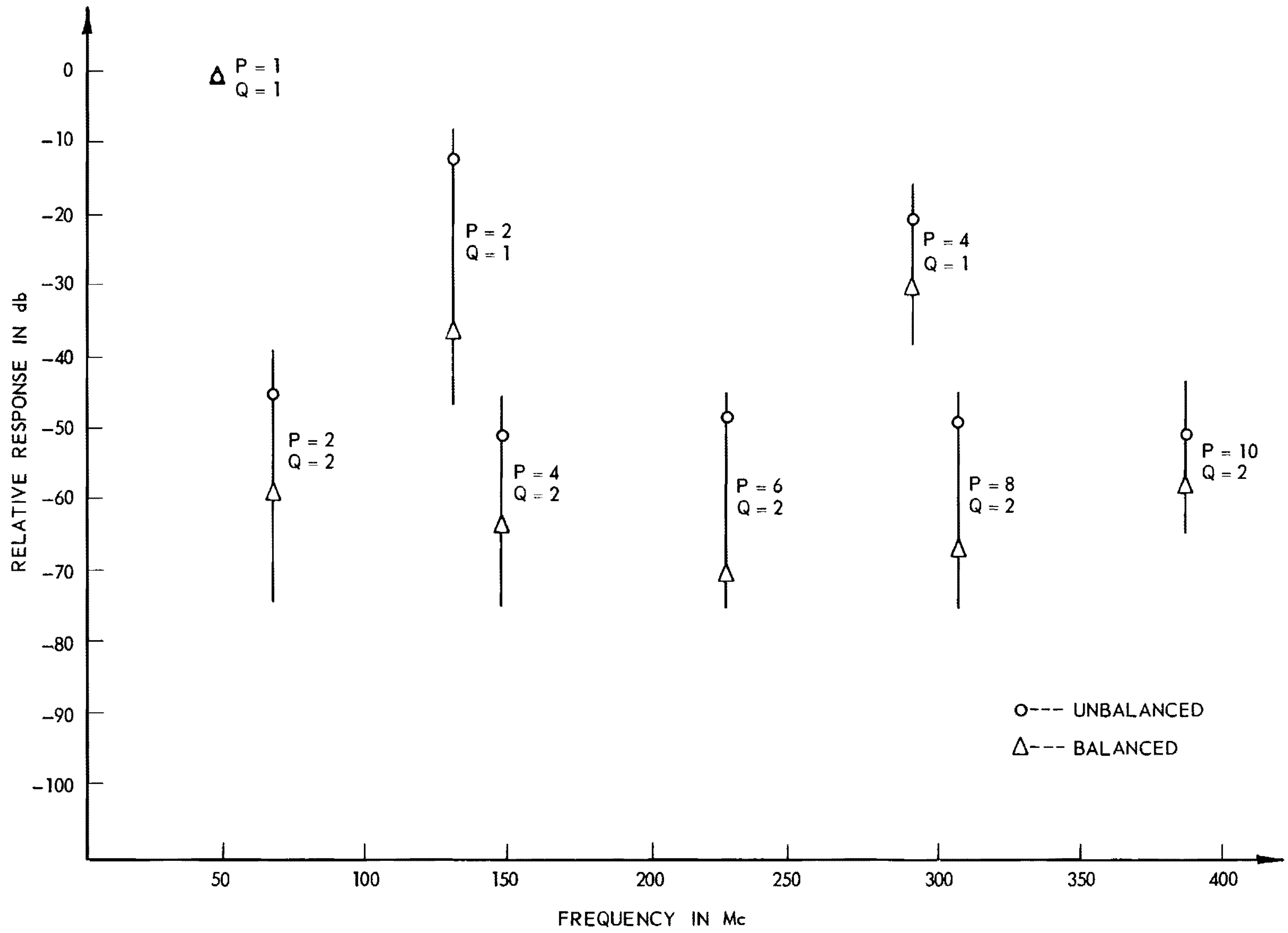



Figure 26. Response Levels for Balanced and Unbalanced Mixers

spurious response rejection over the unbalanced case, but this adjustment is difficult to maintain in practice. An improvement of 20 db is a realistic figure for the cancellation which can be obtained without the need for persistent adjustment of the mixer balancing.

To obtain this cancellation of even order components, it is necessary to match the characteristics of the two mixers carefully so that the amplitudes of the various IF components developed in each mixer are of equal amplitudes and may be cancelled. In addition, the cancellation depends upon the fact that the local oscillator signals are $180^{\circ}$ out of phase and contain no appreciable harmonic components. This condition is not generally met by conventional local oscillators. As a result, the expected cancellation of even order responses is not always obtained in practice. In order to realize the expected cancellation of even order responses, it is necessary to provide two local oscillator signals whose harmonic content is quite low. This harmonic filtering is necessary because the balancing action of the mixer cancels only the internally generated local oscillator harmonics. As a result, even small amplitudes of the harmonics being fed in from the local oscillator can upset the balancing action. This need for harmonic filtering of the local oscillator signal is generally not necessary when using a single ended mixer, since the level of the harmonics generated inside the mixer is usually quite large, with respect to those harmonics contained in the local oscillator signal.

The construction of an oscillator having the necessary phase and amplitude balance as well as the required harmonic filtering is shown in Figure 27. The push-pull line-controlled oscillator in the lower compartment has its

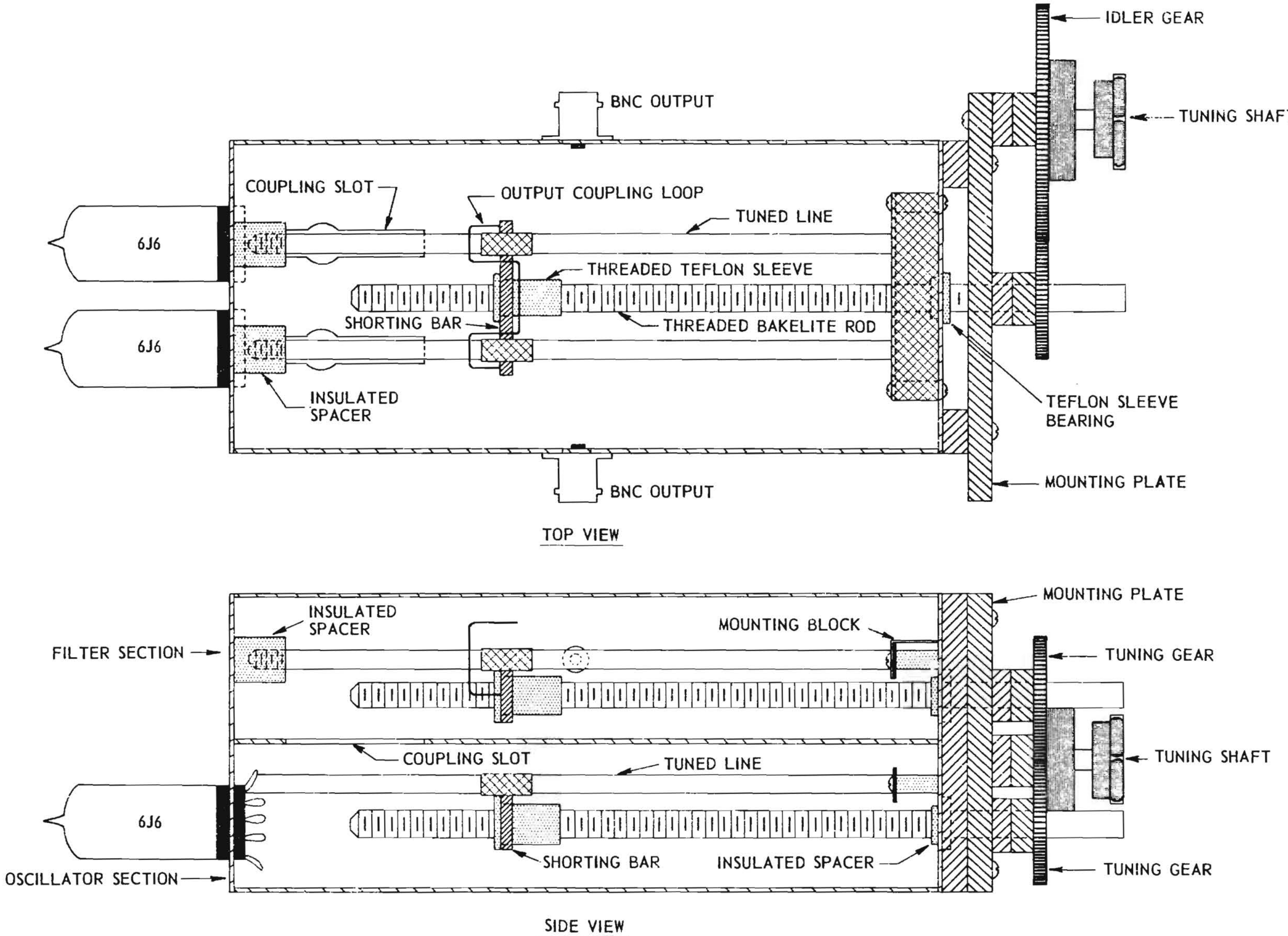



Figure 27. Local Oscillator Assembly

output coupled through two slots in the dividing wall to the tuned-line filter in the upper compartment. The two out-of-phase outputs are obtained by means of a balanced link which is inductively coupled to the tuned-line filter. Tuning is accomplished by means of a threaded bakelite rod which engages a threaded teflon insert on the sliding shorting bar. Since the threaded rod is held in place by retainer rings, rotation of the rod causes the shorting bar to move, changing the frequency to which the line is tuned. The oscillator and filter sections are mechanically coupled through an idler gear to permit tuning of both circuits with one shaft. The photograph of Figure 28 shows an overall view of the local oscillator assembly, and the schematic of Figure 29 gives the details of the oscillator and filter circuitry.

2.3.3.4 Switching Mixers: Another approach to obtaining mixing action is illustrated in Figure 30. In this technique, the mixing is accomplished by periodic operation of a switch in series with the input signal; the period of the switch being that of the local oscillator. The gain of such a switching network to the desired signal has two states, one corresponding to the switch closed and the other corresponding to the switch open. Both these states are linear with respect to the input signal, with the result that no harmonics of the input signal can be generated in either position of the switch. If the transmission of the network is written in the form of Equation (13), where the output is represented as a product of the input signal and the transfer function of the switch, then a Fourier series expansion of the time variable switch transfer function may be made to give the result of Equation (14). Since the switch is operated periodically at the local oscillator rate, the expansion can contain only terms

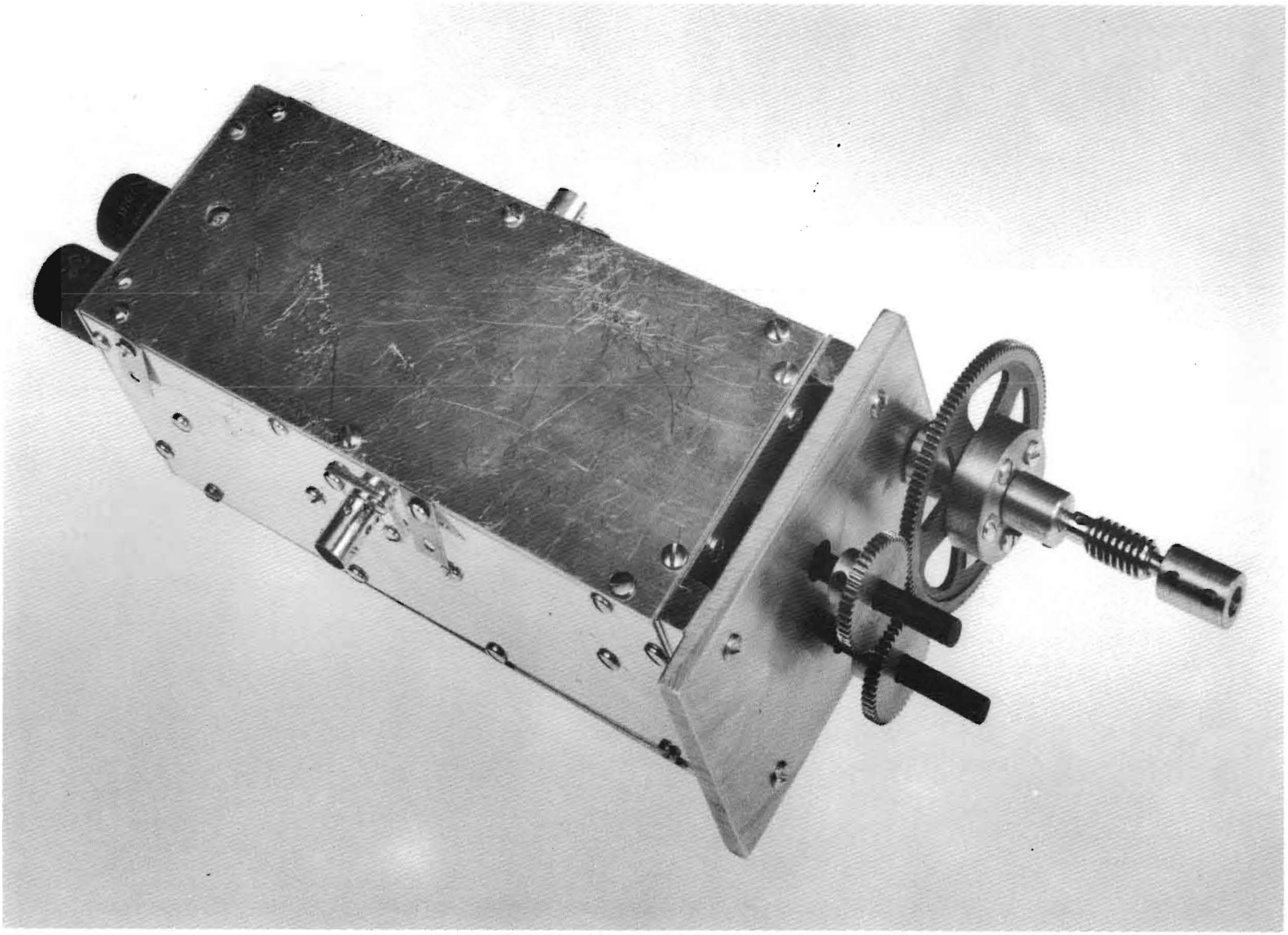


Figure 28. Overall View of Local Oscillator

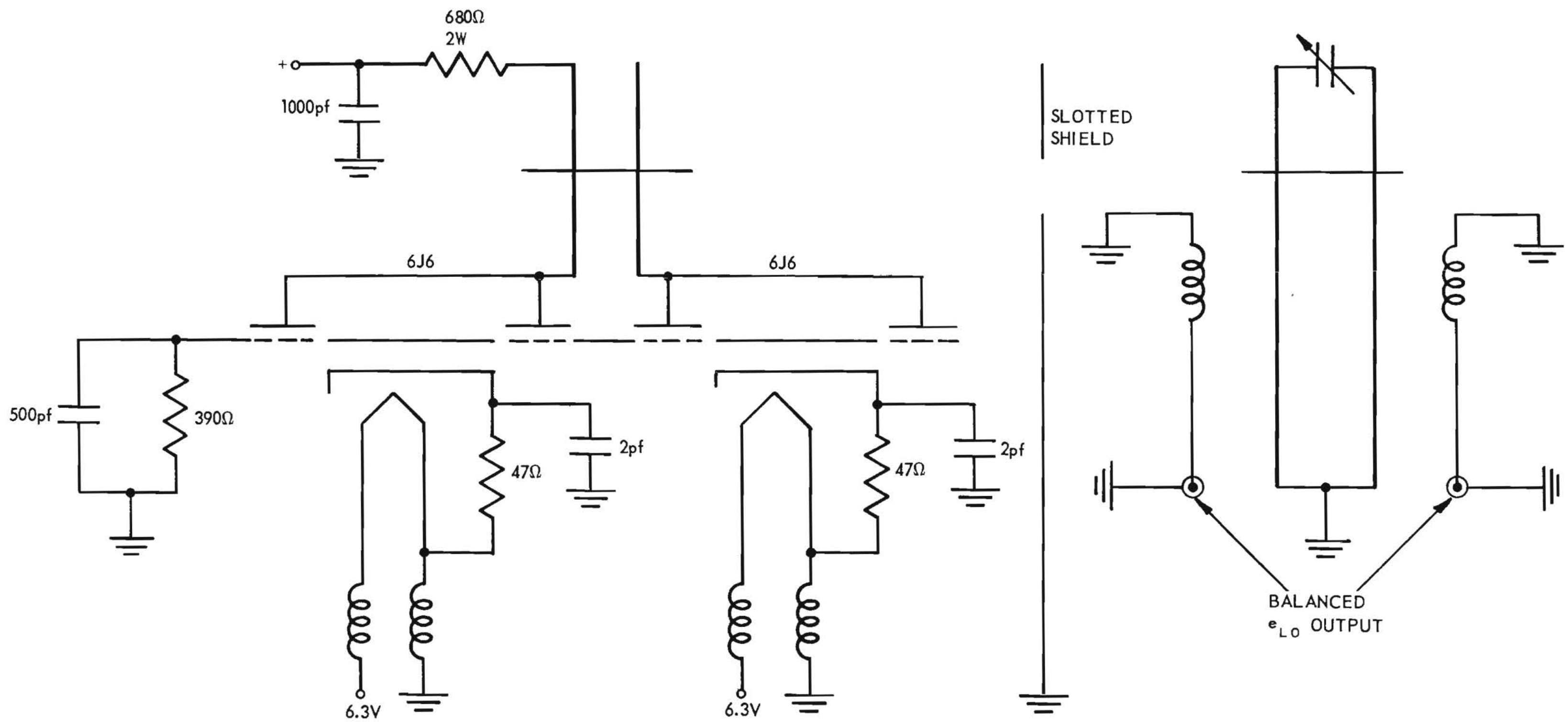
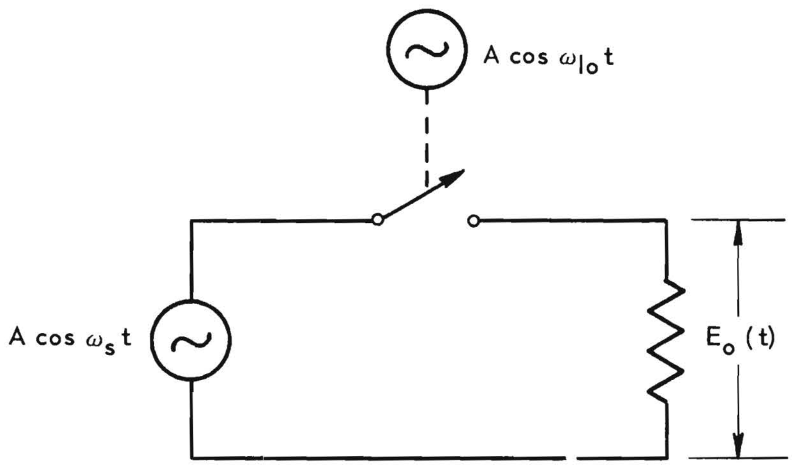



Figure 29. Local Oscillator Schematic




$$E_o(t) = [A \cos \omega_s t][T(t)] \begin{cases} A \cos \omega_s t = \text{Input Voltage} \\ E_o(t) = \text{Output Voltage} \\ T(t) = \text{Switch Transfer Function} \\ \qquad = \sum_{P=0}^{\infty} a_p \cos P\omega_{lo} t \end{cases}$$

$$E_o(t) = \frac{A}{2} \sum_{P=0}^{\infty} a_p \cos (P\omega_{lo} t \pm \omega_s t)$$

Figure 30. Switching Mixer

$$E_o(t) = [A \cos \omega_s t]\, T(t) \text{ -----} \begin{cases} A \cos \omega_s t & = \text{ input voltage} \\ E_o(t) & = \text{ output voltage} \\ T(t) & = \text{ transfer function} \end{cases} \qquad (13)$$

$$E_o(t) = \frac{A}{2} \sum_{P=0}^{\infty} a_n \cos \left(P\omega_{lo} \pm \omega_s\right) t \qquad (14)$$

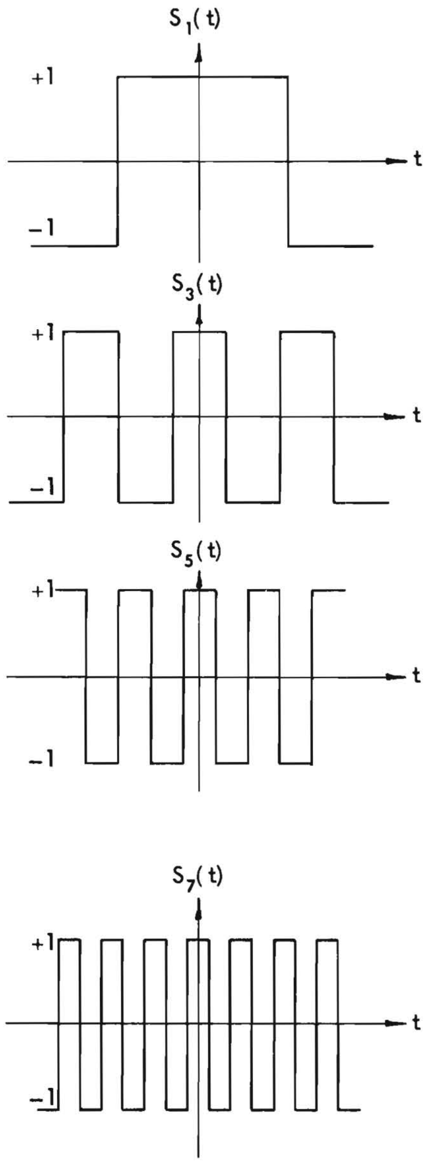
whose frequency is some integral multiple of the local oscillator frequency. Consequently, term by term multiplication of the input signal across the summation sign yields all the output components of the time varying network. Notice that no harmonics of the input signal appear. As a result, the responses of the mixer which are commonly associated with harmonics of the input signal are no longer present. However, there are still those responses remaining which arise from mixing of the input signal with harmonics of the local oscillator. First, the use of a balanced switch will reduce the responses due to even order harmonics so that only the responses arising from odd order harmonics of the local oscillator will be of significant magnitude. In addition, the frequency locations of the remaining responses are far enough away from that of the desired signal that reasonable RF pre-selection can reduce the level of these responses to a very low value. For example, if the desired signal is at a frequency of 50 mc and the local oscillator is at a frequency of 60 mc, resulting in an IF frequency of 10 mc, the response arising from the third harmonic of the local oscillator has a frequency of 170 mc. This frequency spacing from the desired signal amounts to almost two octaves, with the result that this response is readily rejected by the incorporation of a simple RF preselector in front of the mixer.

It is possible to use switching functions other than a simple on-off function so that the harmonic amplitudes associated with the switching function might be drastically reduced. One possibility is the synthesis of a time variable switching function of the form $\cos \omega_{lo} t$. This can be accomplished by forming the desired transfer function as the sum of a series of switching functions of the conventional rectangular form. For instance, a suitable set of functions would look like those in Figure 31. The physical realization of this synthesis might take the form of a number of switches in parallel, each of which is operated periodically at either the local oscillator rate or one of its harmonics. The output of each switch is fed through an attenuator whose loss is set in proportion to the harmonic amplitude associated with that particular harmonic to a common junction at the output, forming the sum of all these switching functions. The over-all transfer of the switching array then closely approximates the desired cosine function.

In the practical realization for any such system, it must be recognized that there is finite switching time associated with the switching element and a nonlinear transfer of the input signal is possible when the switching device is in the transition state. The amount of this nonlinear effect is directly related to the per cent of the total time which the switching spends in the transition state so that for a given transition time, the nonlinear effects of the switch increase directly with the switching rate. As a result, it is necessary to use a high speed switching device or to reduce the frequency at which this switching technique is applied.

In summary then, there are several methods by which the spurious response

$$\cos \omega_{lo} t = C_1 S_1(t) + C_3 S_3(t) + C_5 S_5(t) + - - -$$




$$S_1(t) = \frac{2}{\pi}\left\{\cos \omega_{lo} t - \frac{1}{3}\cos 3\omega_{lo} t + \frac{1}{5}\cos 5\omega_{lo} t - \frac{1}{7}\cos 7\omega_{lo} t + \frac{1}{9}\cos 9\omega_{lo} t - \frac{1}{11}\cos 11\omega_{lo} t + \frac{1}{13}\cos 13\omega_{lo} t + \cdots\right\}$$

$$S_3(t) = \frac{2}{\pi}\left\{\cos 3\omega_{lo} t - \frac{1}{3}\cos 9\omega_{lo} t + \frac{1}{5}\cos 15\omega_{lo} t + \cdots\right\}$$

$$S_5(t) = \frac{2}{\pi}\left\{\cos 5\omega_{lo} t - \frac{1}{3}\cos 15\omega_{lo} t + \frac{1}{5}\cos 25\omega_{lo} t + \cdots\right\}$$

$$S_7(t) = \frac{2}{\pi}\left\{\cos 7\omega_{lo} t - \frac{1}{3}\cos 21\omega_{lo} t + \frac{1}{5}\cos 35\omega_{lo} t + \cdots\right\}$$

BY INSPECTION: $C_1 = 1,\ C_3 = \frac{1}{3},\ C_5 = -\frac{1}{5},\ C_7 = \frac{1}{7}$

Figure 31. Harmonic Switching Functions

rejection of mixers may be improved over that commonly encountered. They are: (1) carefully select an operating point and maintain this point by the use of fixed bias, (2) operate the mixer at the lowest signal level consistent with signal-to-noise ratio and adequate conversion gain, (3) carefully balance the mixers for both desired and local oscillator signals, or (4) operate a mixer in a high speed linear switching mode so as to reduce the nonlinearity presented to the input signals. In every case, provide maximum amount of preselection which can be practically obtained.

2.3.4 Front End Adaptor: A collection of spurious response suppression devices has been assembled into a single unit so as to more readily demonstrate the applicability of each technique to given interference situations. The manner in which this has been accomplished can be seen in the photograph of Figure 32 which shows an internal view of the completed equipment. The physical location of each device is indicated on the photograph. The RF attenuator has been included to prevent overloading of the mixer and IF preamplifier on very strong signals. A total of 50 db of RF attenuation is available which can be selected in 10 db steps by means of push-buttons located on the front panel.

The mixer actually selected for use in the front end adaptor is that shown in the schematic of Figure 33. This mixer is a balanced grounded grid arrangement which rejects spurious responses associated with even harmonics of the local oscillator. Separate control of the DC operating point is provided so that the maximum rejection of responses due to odd harmonics so that local oscillator consistent with adequate gain conversion is obtained.

Two 30 mc IF amplifier stages have been added to the mixer output to

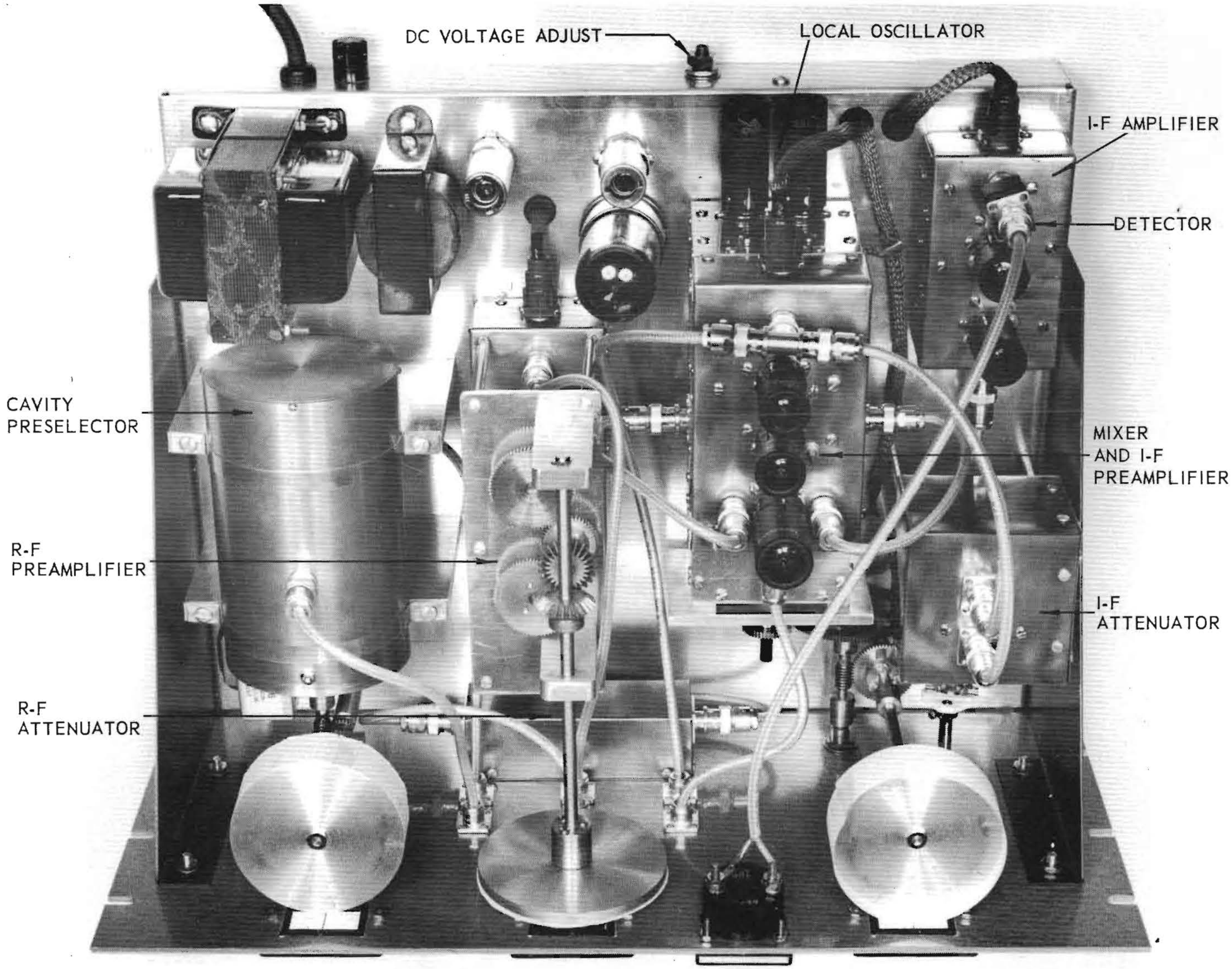



Figure 32. Internal View of Front End Adaptor

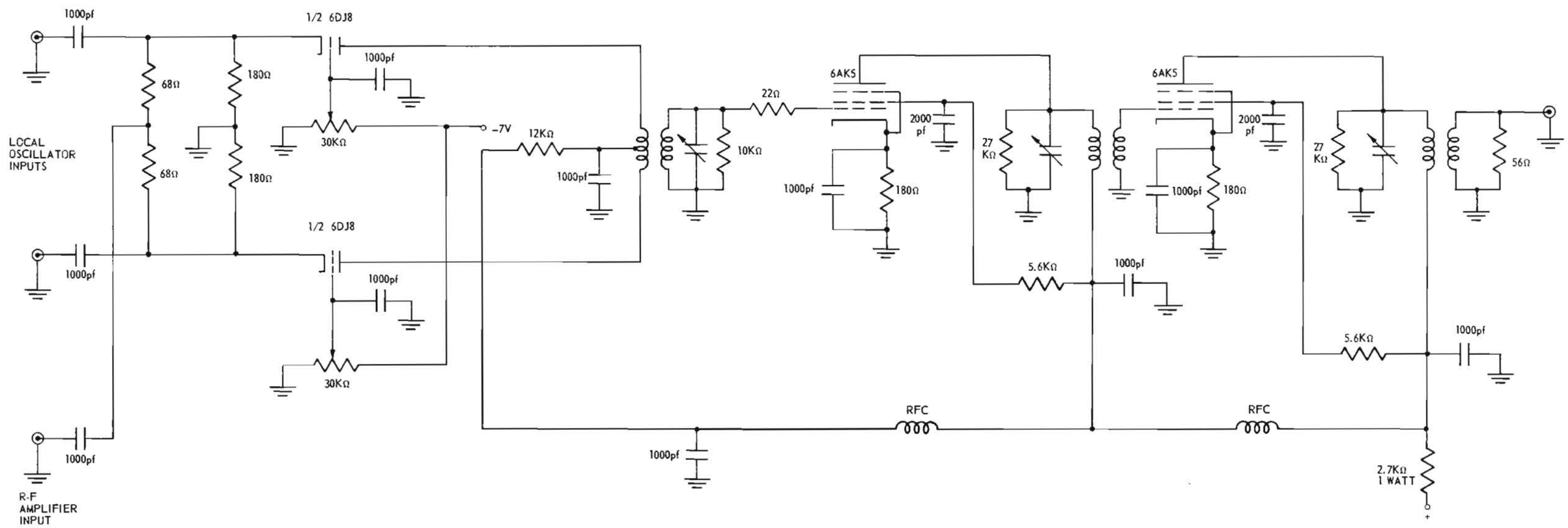



Figure 33. Mixer Schematic

raise the level of the IF signal to a readily usable value. The IF preamplifier output is fed to a front panel output jack and also to an additional two stage 30 mc amplifier which provides a sufficiently large signal to drive a detector and front panel meter. 40 db of switch selected attenuation is provided at the input to this amplifier to permit "on scale" readings of the signal level meter over a wide range of input signal levels. Since no mechanical tracking has been provided between the cavity preselector, the RF amplifier, and the local oscillator, some form of tuning indicator is essential to indicate the correct position of the three tuning controls. The signal level meter at the IF output provides the necessary indication of correct tuning.

A regulated dc power supply has also been included to supply the necessary dc power to each of the suppression devices that need such a power supply, so that 115 volt ac power is all the input required to demonstrate any or all of the devices included in the adaptor. An adjustment is provided at the rear of the chassis for setting the dc voltage at the output of the regulated power supply. This regulated dc voltage should be set at 140 volts. Figure 34 shows the arrangement of the front panel controls for the adaptor. To place the adaptor in operation, the desired combination of devices is selected by proper connection of the jumper cables on the front panel jacks. The tuning controls are then set to approximately the correct positions by means of the calibration on the tuning dials, and are set to exactly the right positions by peaking the reading on the signal level meter. The proper values of RF and IF attenuation are set in to prevent overload conditions and to give an on scale reading of the tuning meter. The adaptor is then ready to operate as desired.

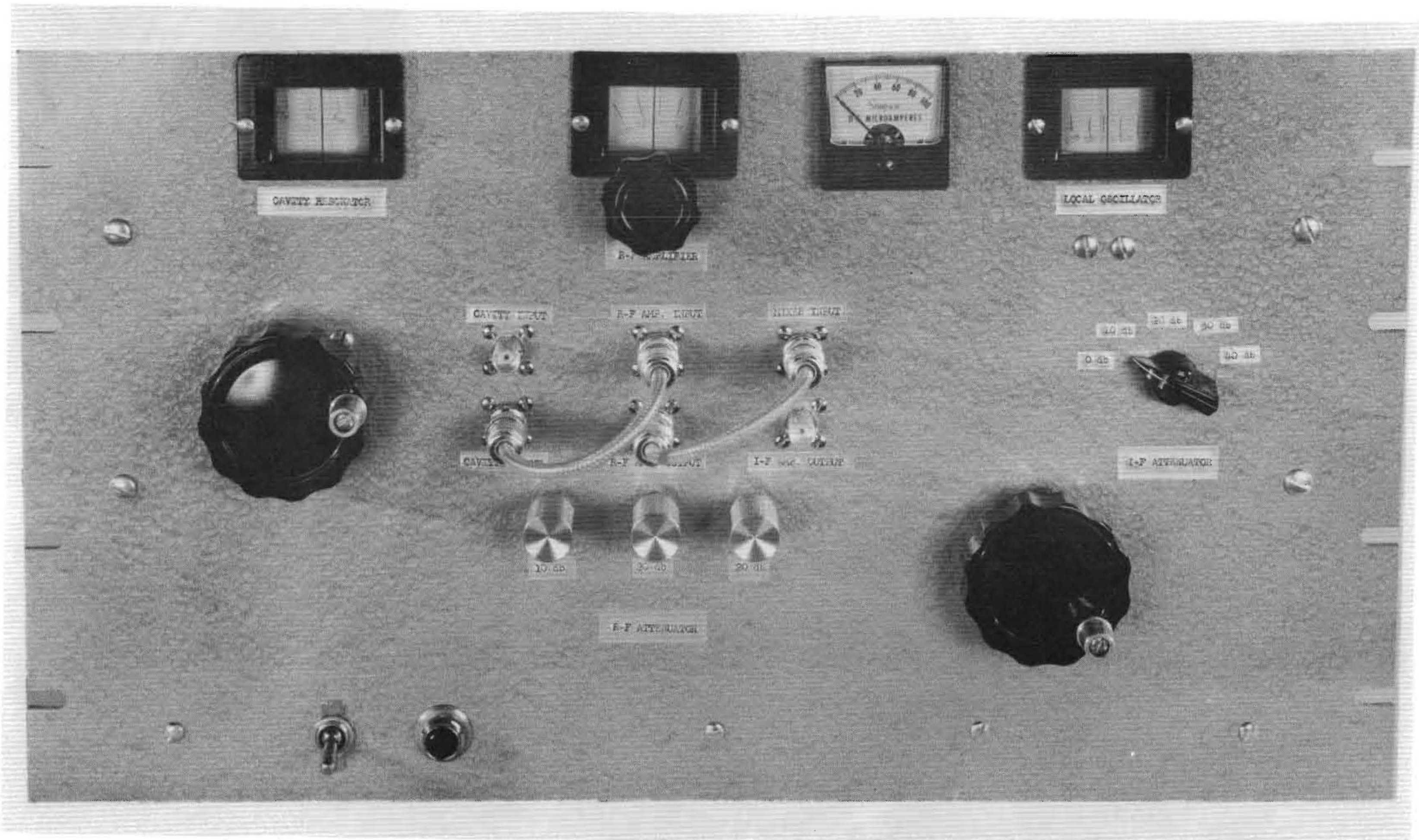



Figure 34. Adaptor Front Panel

## 2.4 Audio Filters for Periodic Signals

At times, periodic interference is present at the input of a communications receiver at a level which, though not of sufficient amplitude to overload the receiver, produces annoying interference in the receiver output. In such an event, an audio filter which will reject this periodic signal is of considerable use in improving the readability of the output audio. Another instance where such a filter can be used to advantage is that which arises when an RF blanker is being used in the front of the receiver to remove high level pulse interference. The blanking action produces periodic amplitude modulation of the desired signal which appears as a tone interference in the audio output of the receiver. An audio filter can remove this interfering tone without appreciably altering the desired signal. One means of accomplishing this filtering is the delay line filter shown in Figure 35.

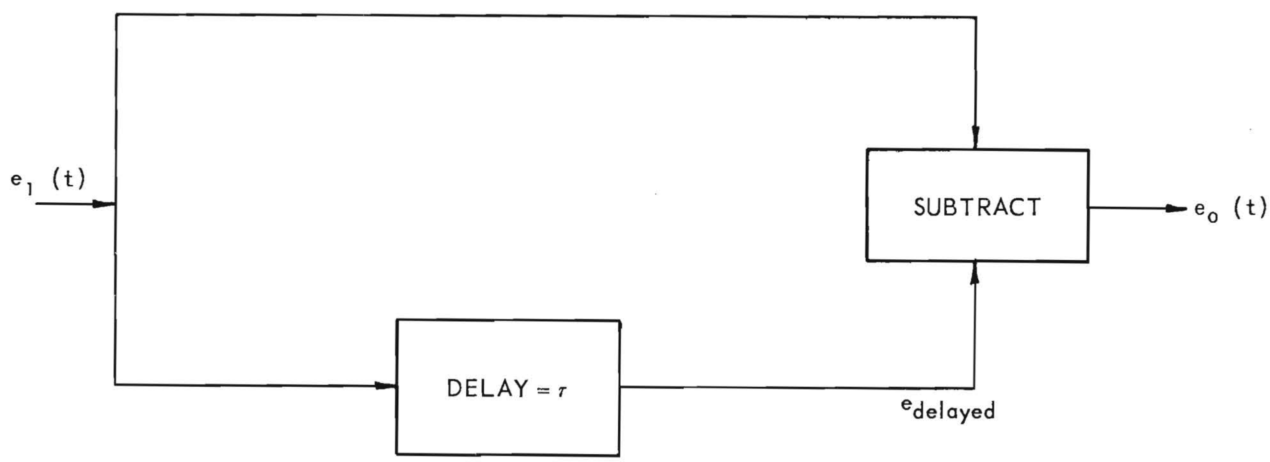



Figure 35. Delay Line Filter

The configuration is a simple cancellation scheme by which the phase shift through the delay line is some multiple of $360^{\circ}$ at the fundamental frequency and at each of the harmonics of the periodic signal. The following analysis illustrates the exact form of the frequency response of such a filter.

Referring to Figure 35, the input signal is:

$$e_{in} = e_1 (t) . \tag{15}$$

Then the delayed signal will be given by:

$$e_{delayed} = e_1 (t - \tau) , \tag{16}$$

and the output signal is:

$$e_o(t) = e_1 (t) - e_1 (t - \tau). \tag{17}$$

Taking the Laplace transform of both sides gives:

$$E_o(s) = E_1(s) - E_1(s)e^{-s\tau} , \tag{18}$$

or:

$$\frac{E_o(s)}{E_1(s)} = 1 - e^{-s\tau} . \tag{19}$$

The steady state response is obtained by letting:

$$s = j\omega . \tag{20}$$

Then:

$$\text{Gain } (j\omega) = \frac{E_o(j\omega)}{E_1(j\omega)} = 1 - e^{-j\omega\tau} . \tag{21}$$

The zeros of this gain function occur for:

$$e^{j\omega\tau} = \cos \omega\tau + j \sin \omega\tau = 1 + j0 , \tag{22}$$

or:

$$\omega = \frac{2N\pi}{\tau} \tag{23}$$

These are the values of $\omega$ which the fundamental and harmonic frequencies of the periodic interference must have if it is to be rejected. A sketch of the gain function of equation (21) is shown in Figure 36. The positions of the zeros of this gain function can be moved so as to reject a periodic signal by varying the value of $\tau$ until $\tau$ is equal to the period of fundamental of the periodic signal, i.e.,

$$\tau = \frac{1}{f_{(\text{fund.})}} . \tag{24}$$

An audio filter mechanizing the block diagram of Figure 35 has been constructed and its ability to reject periodic signals in the manner indicated by the frequency response curves of Figure 36 has been demonstrated.

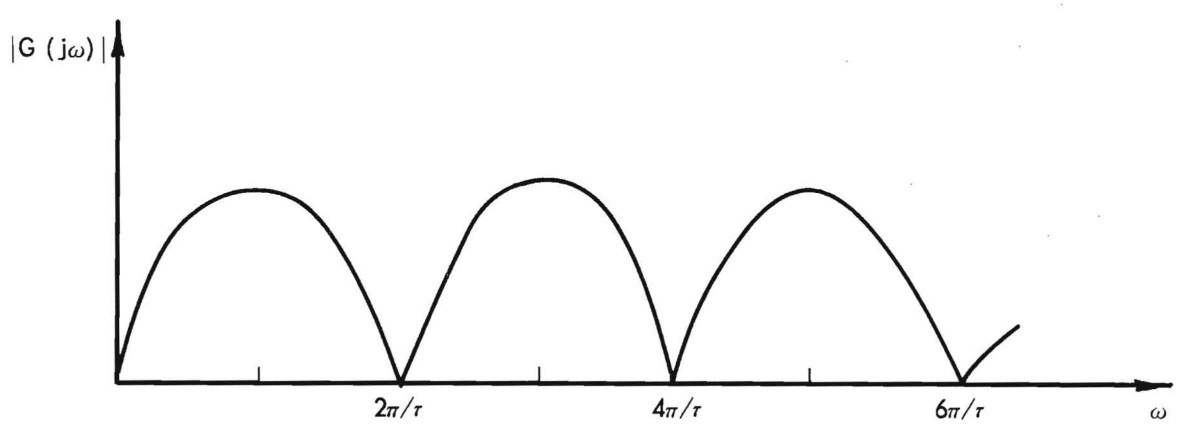



Figure 36. Response of Delay Line Filter

The most difficult part in the design of such a filter for operation at audio frequencies is the construction of the variable delay line. The manner in which this delay is obtained is shown in Figure 37. The amplitude of the signal to be delayed is sampled at a sufficiently high rate to accommodate the required bandwidth. For a typical communications receiver this bandwidth is about 5 kc. These samples are then used to pulse position modulate the output of the reference oscillator. This operation codes the signal information into the time position of the signal pulse with respect to the reference pulse, so that a chain of delay multivibrators can be used to obtain the desired delay. Subsequent demodulation of the output of the multivibrator delay chain yields a delayed reproduction of the input signal. Adjustment of the delay is obtained by varying the delay time in the multivibrators. The pulse position modulator is a Schmidt trigger circuit whose input consists of the sum of the modulating signal and a sawtooth voltage which is triggered at the reference pulse rate. The Schmidt trigger is set to fire when its input signal crosses through zero voltage, which occurs whenever the sawtooth voltage is equal but opposite in phase to the modulating signal.

The time with respect to the start of the sawtooth and hence to the reference pulse, at which this zero crossing occurs is a linear function of the amplitude of the modulating signal. As a result, the time position of the output pulse from the Schmidt trigger circuit varies linearly with the amplitude of the modulating signal. The diagram of Figure 38 illustrates this modulation arrangement.

The basic delay section used in the delay chain is shown in Figure 39. This delay section is a one shot multivibrator whose delay time can be

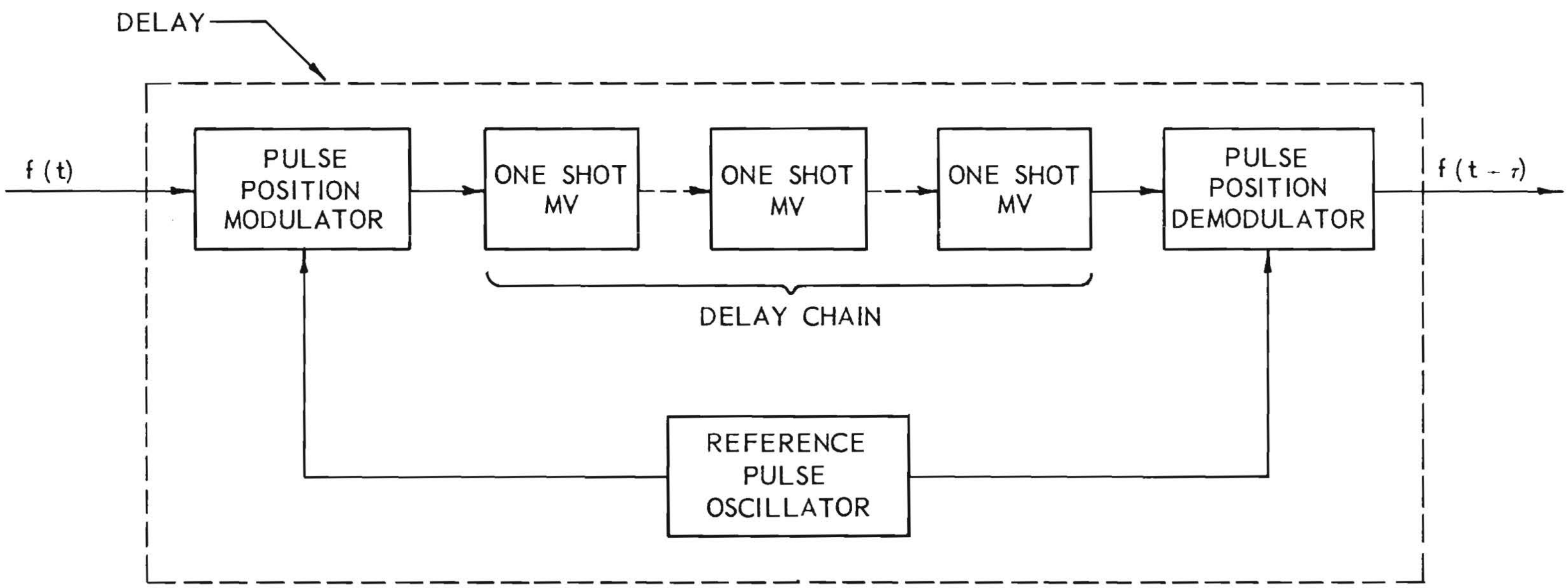



Figure 37. Varable Delay Line

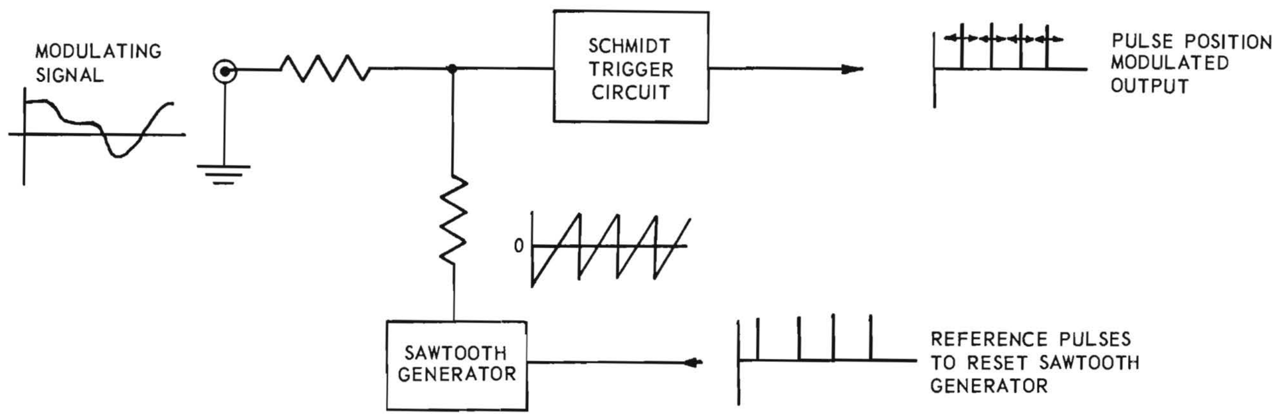



Figure 38. Pulse Position Modulator

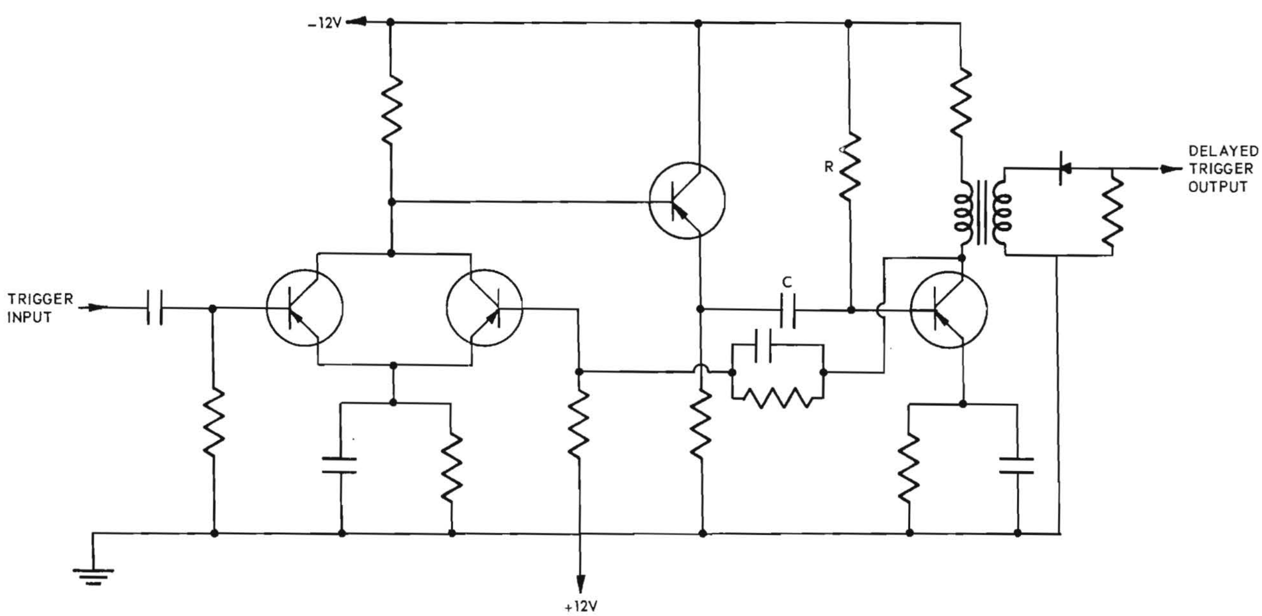



Figure 39. Typical Delay Section Multivibrator

adjusted by means of a resistor (R), as indicated in the figure. Parallel input triggering is used to isolate the trigger source from the remainder of the circuit. Interstage coupling between the two halves of the multivibrator is by means of an emitter follower. This emitter follower provides a low impedance charging path for the timing capacitor, C , so that the recovery time of the multivibrator is quite short, permitting the use of the multivibrator at high duty ratios.

These delay sections have been assembled on printed circuit cards with three multivibrators on each card so that with eight cards, a total delay of approximately 1200 microseconds can be obtained over the entire delay chain. The output of this delay chain is fed to a demodulator where the original modulation is recovered by causing the output developed pulse from the delay chain to sample a sawtooth waveform which has been initiated by the reference pulse. The amplitude of a particular sample of the sawtooth waveform is a linear function of the time delay of the multivibrator output with respect to the reference pulse. This sample value is stored on a holding capacitor so that the output voltage across the capacitor is a reproduction of the original signal delayed by an amount equal to the delay in the multivibrator chain. A block diagram illustrating this demodulation technique is shown in Figure 40.

An overall view of the completed audio filter, as shown in Figure 41, illustrates the construction techniques employed.

Figure 42 gives a view of the front panel controls. These controls permit selection of any time delay between 500 and 1200 microseconds, thus varying the frequency spacing of the notches in the overall filter response

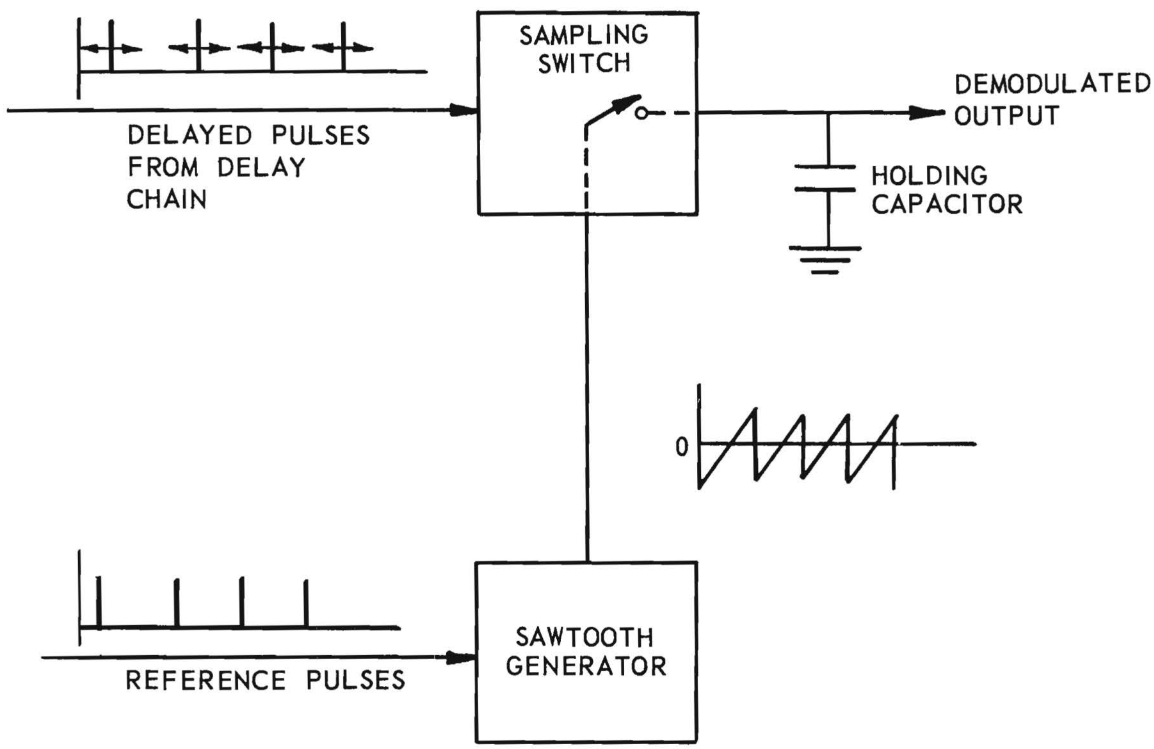



Figure 40. Pulse Position Demodulator

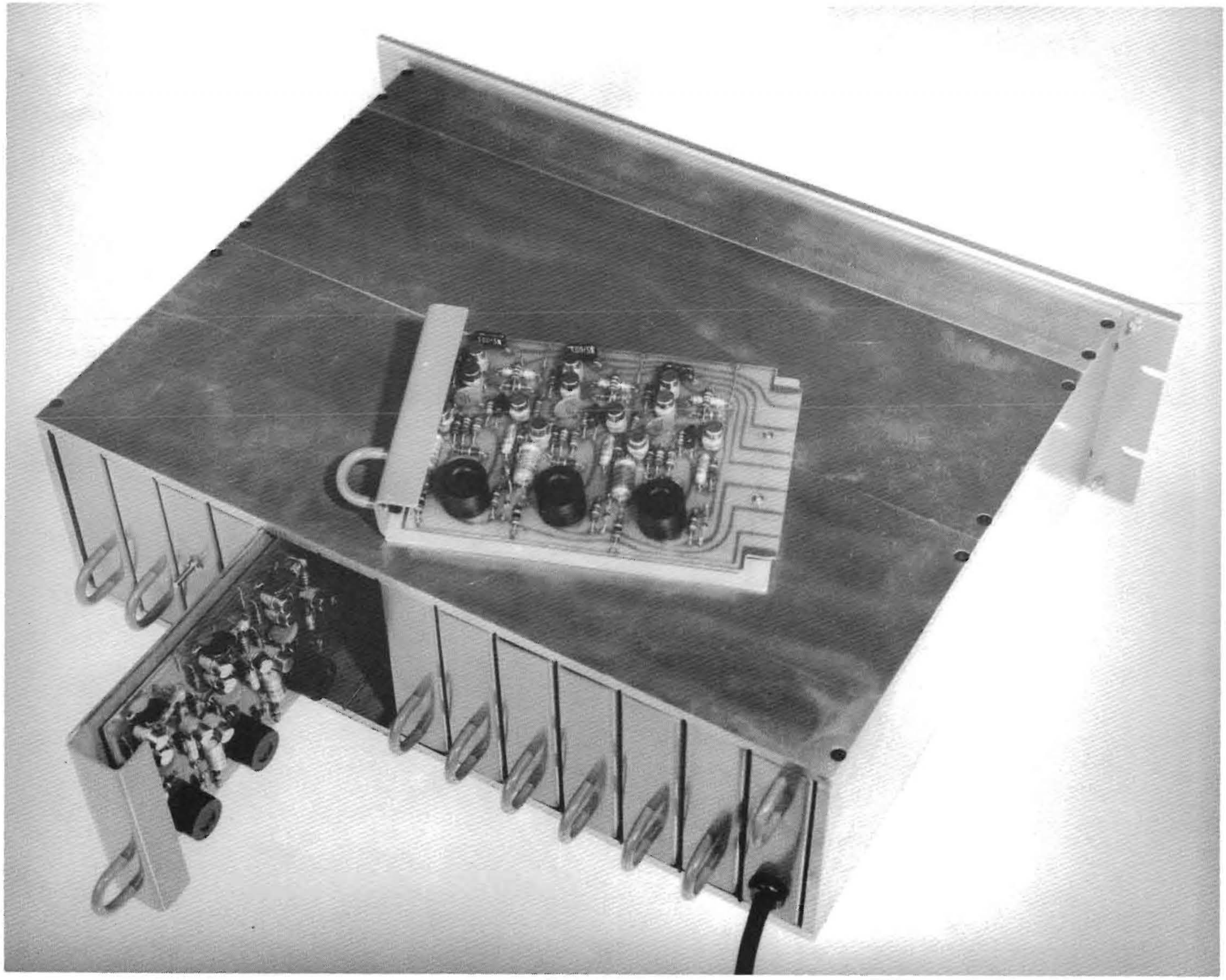


Figure 41. Overall View of Delay Line Filter

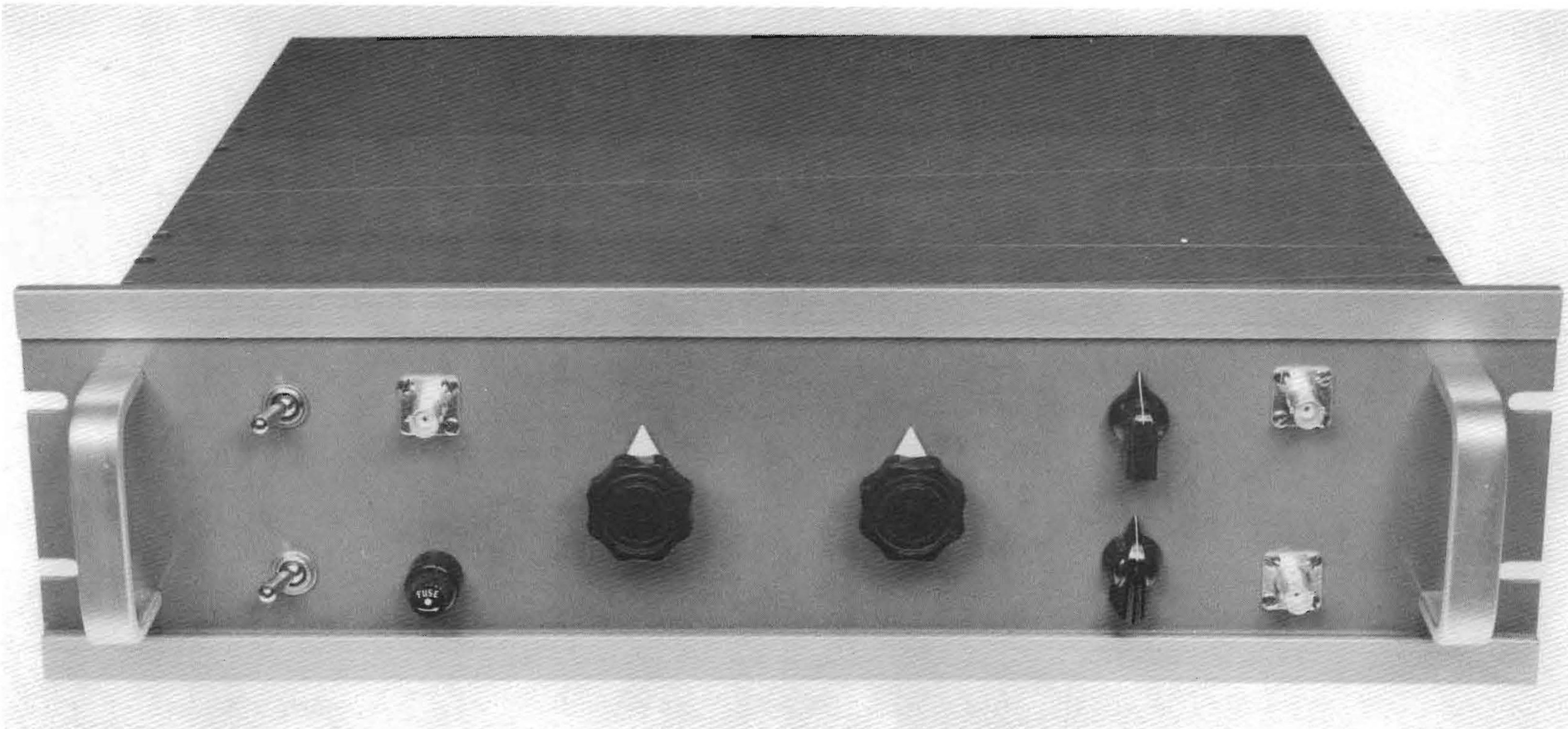



Figure 42. Front Panel of Delay Line Filter

between 2000 and 833 cps. An adjustment is also provided for positioning the reference pulse so that it does not occur in the middle of a modulation cycle. Since, for good cancellation it is necessary to provide accurate control of the phase in both channels, no filtering is performed until after the subtraction process is complete. This eliminates the problem of matching the filter characteristics in the two channels. Using this filter, periodic signals may be rejected by as much as 30 db. A complete schematic of this filter is given in the Appendix.

### 2.5 Tunnel Diode Mixers

In many instances a measurements antenna may be remotely located from a field intensity meter. In such a circumstance, excessive cable losses may attenuate signals below the noise level of the field intensity meter. The location of the mixer directly at the antenna terminals may overcome this difficulty since the cable attenuation may be considerably lower at the IF frequency than at the signal frequency.

Conventional diode mixers have an inherent conversion loss so that the use of such mixers is an advantage only in the circumstance where the conversion loss does not exceed the reduction in attenuation in the cable gained by operating the cable at the IF frequency. For this reason, a tunnel diode mixer offers promise, since its conversion gain may be considerably greater than that obtained with a conventional diode mixer.

This increased conversion gain is obtained if the time varying conductance of the tunnel diode is negative over a portion of the local oscillator cycle. If the negative conductance is large enough, the conversion gain of the mixer can be greater than one.

Figure 43 illustrates one possible configuration of a tunnel diode mixer. The conversion gain is a function of the local oscillator amplitude since a certain minimum local oscillator signal is necessary to carry the operating point into the negative resistance portion of the tunnel diode characteristic. When the proper amplitude of local oscillator signal is applied, the conversion gain can exceed that of a conventional diode mixer by a considerable amount. The high pass filter at the input prevents the IF signal from being dissipated in the output resistance of the signal source, while the low pass filter at the IF output prevents the incoming signal from being dissipated in the IF impedance.

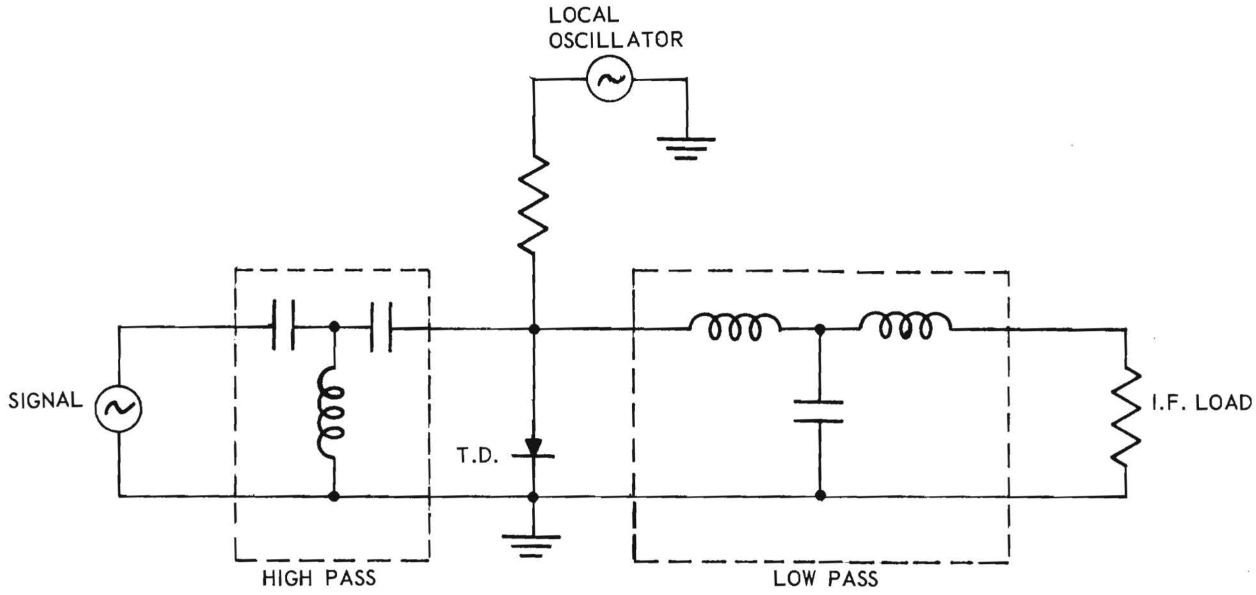



Figure 43. Tunnel Diode Mixer

The way in which the conversion gain of the tunnel diode mixer varies with local oscillator amplitude can be explained by an examination of the relative excursions of the local oscillator and desired signals on the non-linear volt-ampere characteristics of the tunnel diode such as that shown in Figure 44.

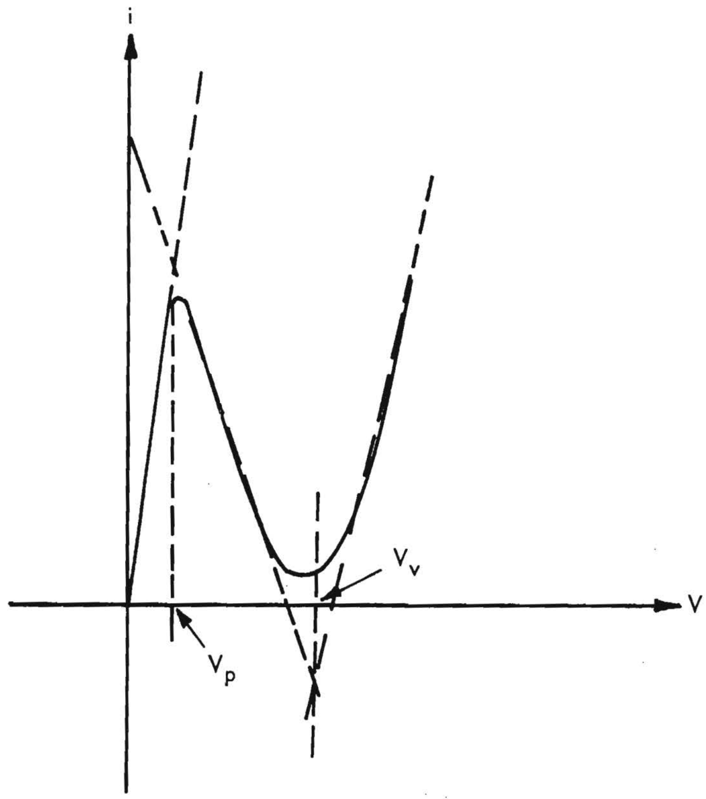



Figure 44. Tunnel Diode v-i Characteristics

By representing the volt-ampere characteristic of the tunnel diode by a series of straight line approximations, the operation of the circuit may be broken up into several distinct regions. The operation of the circuit is linear with respect to the desired signal in each of these regions. For the case where the input signal consists of a large local oscillator voltage added to the small desired signal, the effect of the local oscillator can be thought of as simply controlling the gain of the network to the small desired signal, by switching the operation of the circuit from one of the linear states to another. The technique of the analysis for the tunnel diode mixer then, is to expand the gain function of the circuit, as seen by the small signal, as a Fourier series whose fundamental period is that of the local oscillator. Then by multiplying each term in the expansion by the desired input signal and collecting all terms of the same frequency, the spectrum of the output signal can be obtained. The amplitudes of the various frequency components can then be determined from the values of the coefficients in the Fourier series expansion of the gain function. The ratio of the coefficient of the resulting term at the IF to the amplitude of the input signal is the conversion gain. If the local oscillator and desired signals respectively are given by

$$\begin{aligned} V_{\ell o} &= A \cos \omega_{\ell o} t \text{ , and} \\ i_d &= I_d \cos \omega_d t \text{ ,} \end{aligned} \tag{25}$$

the current in the load is simply the input signal multiplied by the time varying gain of the tunnel diode, i.e.,

$$I_L = (I_d \cos \omega_d t) \sum_{n=0}^{\infty} a_n \cos (n w_{\ell o} t + \phi_n) \ . \tag{26}$$

The intermediate frequency corresponds to n = 1, so that

$$I_{L_{(IF)}} = \tfrac{1}{2} I_d a_1 \cos (\omega_{\ell o} - \omega_d) t \ . \tag{27}$$

The magnitude of the conversion gain is then given by

$$G_{(conv.)} = \frac{I_{L(IF)}}{I_d} = \frac{a_1}{2} \tag{28}$$

If the local oscillator amplitude, A, is less than $V_p$ , the gain variation presented to the desired signal is that shown in Figure 45.

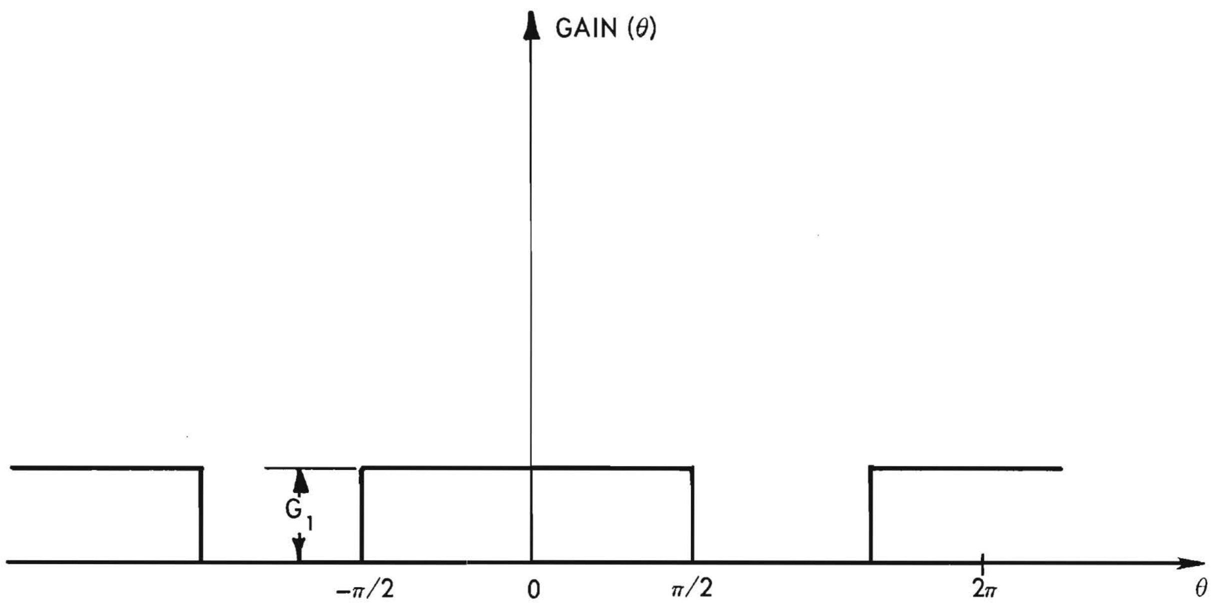



Figure 45. Gain Function for $A \leq V_p$

From Figure 45:

$$a_1 = \frac{1}{\pi}\int_{-\pi}^{\pi} G(\theta)\cos\theta d\theta = \frac{2}{\pi}\int_0^{\pi/2} G_1 \cos\theta d\theta = \frac{2}{\pi} G_1 \sin\theta \Big]_0^{\pi/2} = \frac{2G_1}{\pi} . \tag{29}$$

As the value of A is increased so that

$$V_p \leq A \leq V_v ,$$

the gain function changes to that shown in Figure 46.

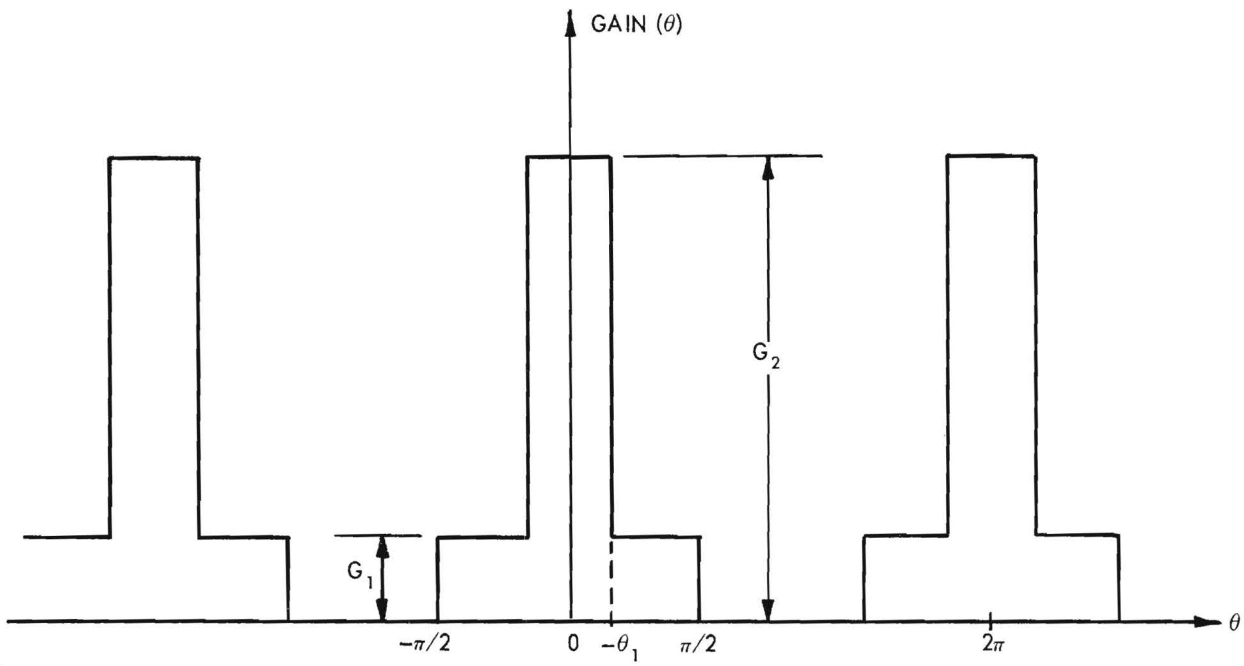



Figure 46. Gain Function for $V_p \leq A \leq V_v$

In this figure, $\Theta_1 = \cos^{-1}\left(\frac{V_p}{A}\right)$.

For this gain function the value of $a_1$ is given as before by

$$a_1 = \frac{1}{\pi}\int_{-\pi}^{\pi} G(\Theta) \cos \Theta \, dt = \frac{2}{\pi}\int_{0}^{\Theta_1} G_2 \cos \Theta \, d\Theta + \frac{2}{\pi}\int_{\Theta_1}^{\pi/2} G_1 \cos \Theta \, d\Theta ,$$

$$= \frac{2}{\pi}\left\{ G_1 + (G_2 - G_1) \sin \Theta_1 \right\}$$

$$= \frac{2}{\pi}\left\{ G_1 + (G_2 - G_1)\sqrt{1 - (V_p{}^2/A^2)} \right\}. \tag{30}$$

If the amplitude of the local oscillator signal is further increased so that $A > V_v$ , the gain function is also changed to appear as shown in Figure 47.

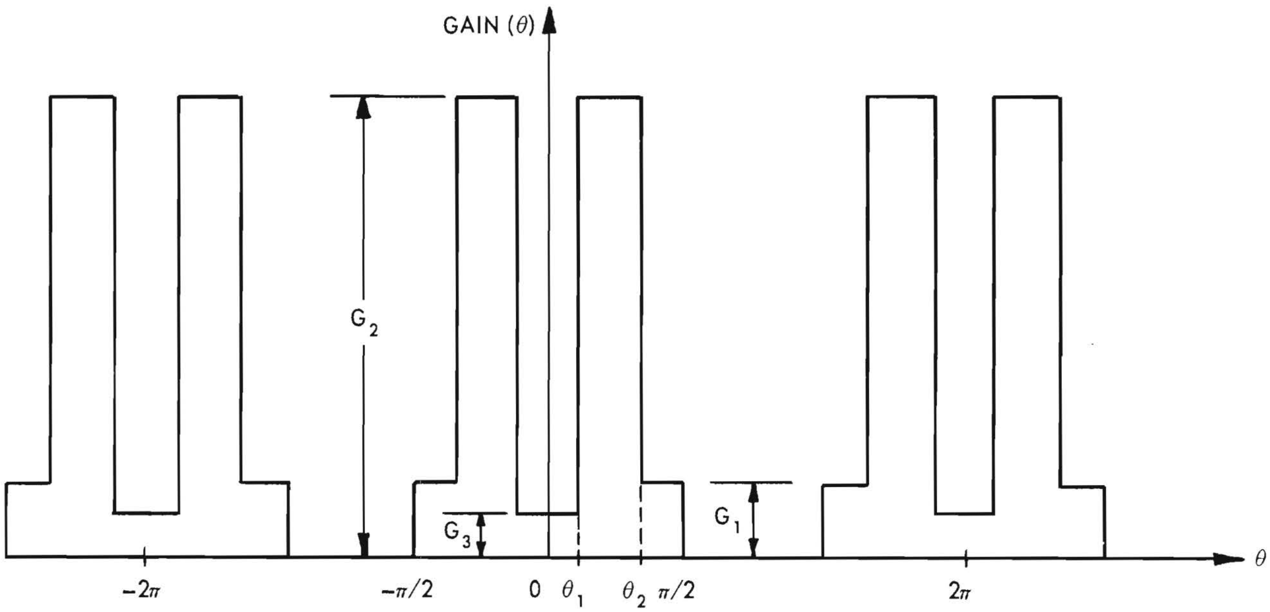



Figure 47. Gain Function $A > V_v$

As before,

$$a_1 = \frac{2}{\pi}\left\{\int_0^{\Theta_1} G_3 \cos\Theta \, d\Theta + \int_{\Theta_1}^{\Theta_2} G_2 \cos\Theta \, d\Theta + \int_{\Theta_2}^{\pi/2} G_1 \cos\Theta \, d\Theta\right\}$$

$$a_1 = \frac{2}{\pi}\left\{ G_3 \sin\Theta_1 + G_2 (\sin\Theta_2 - \sin\Theta_1) + G_1 (1 - \sin\Theta_2)\right\}$$

$$a_1 = \frac{2}{\pi}\left\{ (G_3 - G_2)\sin\Theta_1 + (G_2 - G_1)\sin\Theta_2 + G_1 \right\}$$

$$\Theta_1 = \cos^{-1}(V_v/A) \longrightarrow \sin\Theta_1 = \sqrt{1 - (V_v^{\,2}/A^2)}$$

$$\Theta_2 = \cos^{-1}(V_p/A) \longrightarrow \sin\Theta_2 = \sqrt{1-(V_p^{\,2}/A^2)}$$

so that

$$a_1 = \frac{2}{\pi}\left\{(G_3 - G_2)\sqrt{1-(V_v^{\,2}/A^2)} + (G_2 - G_1)\sqrt{1-(V_p^{\,2}/A^2)} + G_1\right\} \quad (32)$$

Comparison of this expression with the terms for the gain in the other two regions shows that this one expression will hold over the whole range of local oscillator voltage if only the real part of the expression is used, i.e.,

$$\underset{(\text{conv.})}{\text{Gain}} = \frac{1}{\pi} R_e \left\{G_1 + (G_2 - G_1)\sqrt{1-(V_p^{\,2}/A^2)} + (G_3 - G_2)\sqrt{1-(V_v^{\,2}/A^2)}\right\} \quad (33)$$

Now from the circuit shown in Figure 48, the gains $G_1$ , $G_2$, and $G_3$ can be calculated using equation (34).

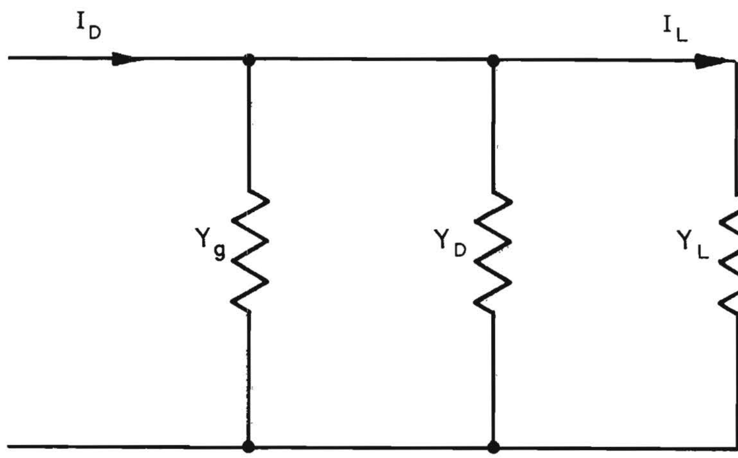



Figure 48. Circuit Used for Gain Computations

Here, $Y_g$ is the generator conductance,

$Y_D$ is the tunnel diode conductance,

$Y_L$ is the load conductance

and the gain, G , is

$$G = \frac{I_L}{I_D} = \frac{Y_L}{Y_g+Y_D+Y_L} \quad . \tag{34}$$

A tunnel diode mixer was constructed in the form shown in Figure 43 and the conversion gain as a function of the amplitude of the local oscillator signal was determined. The results of these measurements are shown in Figure 49, by the curve labeled "experimental". The conversion gain was then calculated using equation (33). The values used for the various quantities in this equation were $V_p = 0.15$ volts, $V_v = 0.30$ volts, $G_1 = 0.047$, $G_2 = 1.0$, and $G_3 = 0.1$. The results of this calculation are also plotted in Figure 43. Considering the fact that only the average data supplied by the tunnel diode manufacturer was available, the correlation between the experimental and calculated values for conversion gain is quite satisfactory.

## 2.6 Modifications to the R-361 UHF Receiver

During a series of tests made on the model R-361 UHF receiver to determine the ability of this receiver to suppress pulse interference, it was found that the internal noise limiter circuit was particularly ineffective in a pulse interference situation. An investigation was made to determine the principles of the noise limiter in the R-361 receiver, the cause of the difficulty encountered during these tests, and means for restoring the effectiveness of the noise limiter circuit.

2.6.1 Operation of The Noise Limiter: The operation of the noise limiter circuit can be explained with the aid of the circuit diagram of Figure 50 which shows the last IF amplifier, the second detector, and the noise

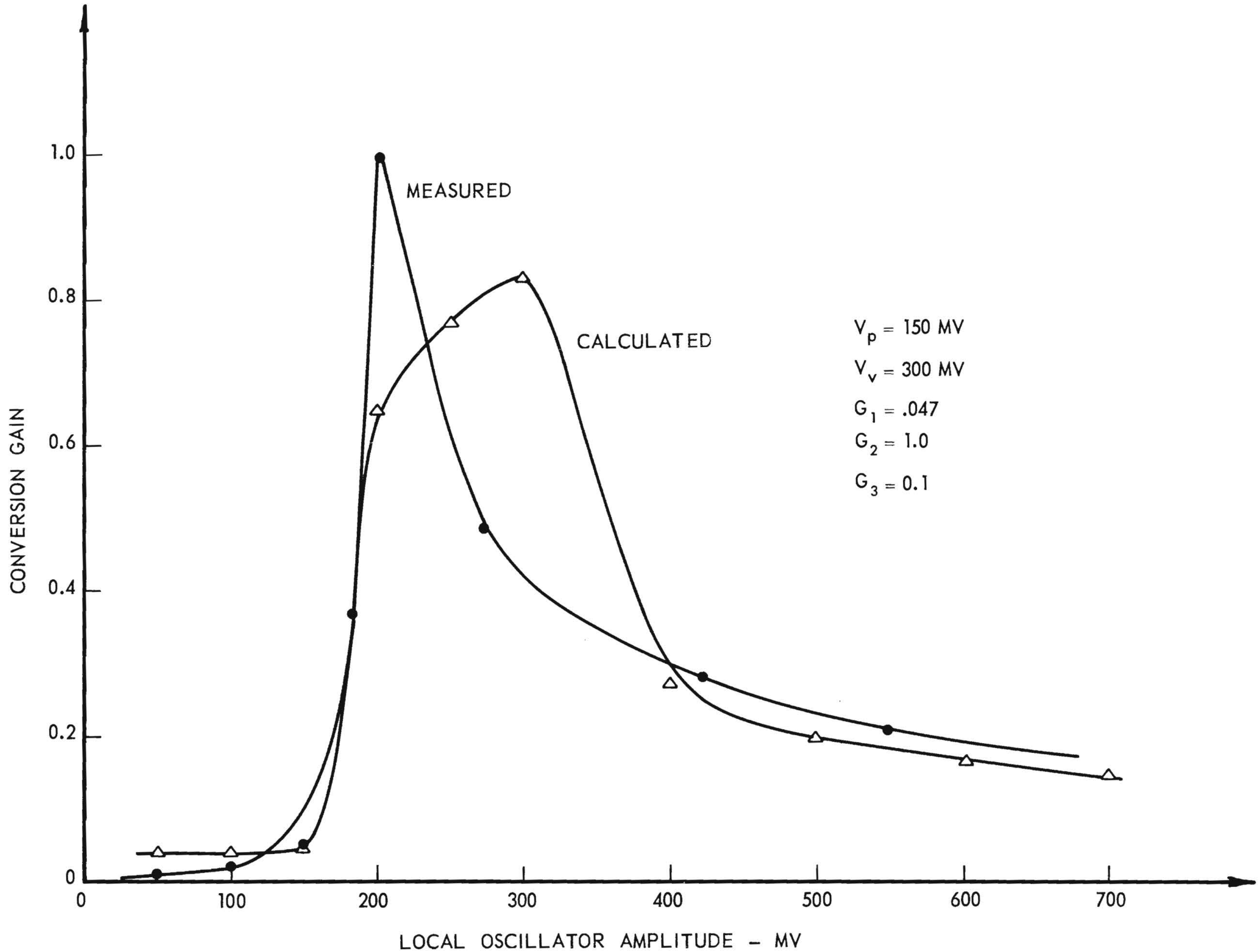



Figure 49. Conversion Gain versus Local Oscillator Amplitude

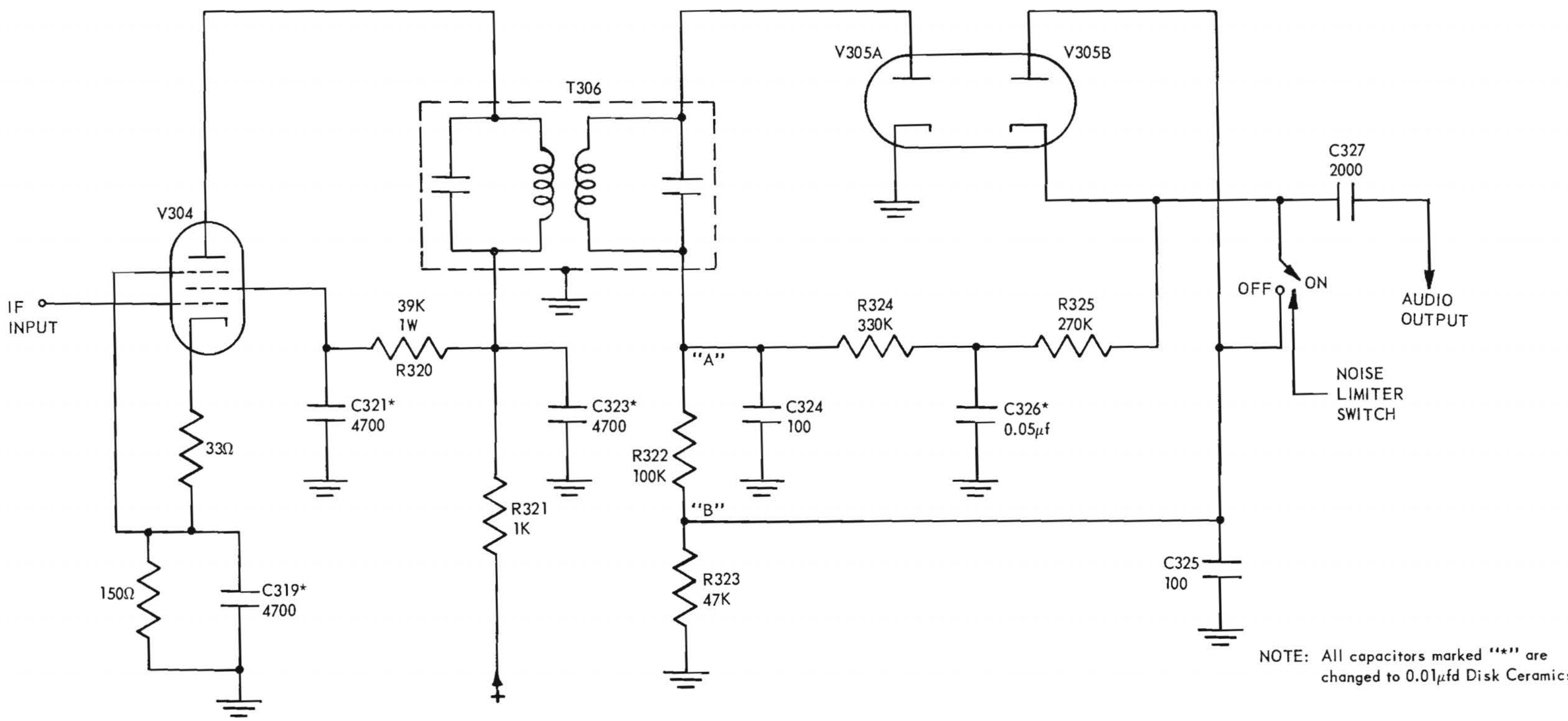



Figure 50. Second Detector and Noise Limiter R361 Receiver

limiter. In this circuit, the amplitude modulated IF signal is rectified by the action of diode V305A causing an audio signal proportional to the envelope of the IF signal to appear between point "A" and ground. A negative dc voltage proportional to the amplitude of the IF carrier component also appears at point "A". This dc component is applied to the cathode of diode V305B through a low pass filter consisting of R324 and C326. The effect of this filter is to prevent any of the audio signal at point "A" from reaching the cathode of V305B while passing the dc voltage at point "A" unchanged to the cathode of V305B. The plate of V305B is tied directly to point "B". The voltage at this point contains the same dc and audio components of the demodulated IF signal as that at point "A", but the magnitude of the signal at "B" is only one third as big as that at "A" because of the voltage divider action of R322 and R323. As a result, V305B has its cathode at a more negative voltage than its plate and is in a conducting state. This permits the audio signal at the plate of V305B to appear at the cathode where it is passed from this point through the audio coupling capacitor C327 to the audio amplifier. Notice that even at 100% modulation, of the desired AM signal, the peak negative swing of the signal at the plate of V305B is still not large enough to reverse bias this diode so that V305B remains in conduction continuously. However, the sudden appearance of a large pulse at point "A", causes a negative pulse of one third this magnitude to appear at the plate of V305B, but the low pass filter composed of R324 and C326 does not permit this pulse to appear at the cathode of V305B. This negative pulse at the plate causes the diode to stop conduction whenever the amplitude of the pulse is larger than three times the carrier level of the AM signal. This disconnect action prevents the interfering pulse from feeding through into the audio channel.

2.6.2 Source of the Difficulty: The disconnect action of the noise limiter diode just described should result in a very effective pulse interference suppressor. However, tests made to determine effectiveness of the noise limiter have shown that very little if any difference can be detected whether the noise limiter is in or out of the circuit. An inspection of the receiver showed that the fault lay in the method of construction rather than in poor circuit design. The close physical location of the components for the audio system input, the noise limiter and the plate and screen decoupling resistors and bypass capacitors for the last IF amplifier, V304, were found to be responsible for the lack of effectiveness of the noise limiter. In particular, the presence of a large pulse signal in the IF amplifier causes a demodulated pulse component of plate current to flow in the last IF amplifier, V304. This pulse of current is drawn through R320 and R321. These resistors are located on the same terminal board as C321, which is the coupling capacitor to the input to the audio amplifier. The stray coupling between R320 and R321 and C321 is sufficient to produce a large audio output whenever a large pulse signal is present, bypassing the audio around the noise limiter. Stray coupling also exists between R320 and R321 and all the components of the noise limiter circuit, making it impossible for normal noise limiter action to take place.

2.6.3 Remedial Action Taken: In an attempt to overcome these difficulties, a relocation of R320 and R321 was tried. This gave some reduction in the stray coupling but the high gain of the audio amplifier in the receiver caused annoying pickup of the pulse signals from the leads to components as well as from the components themselves. Next, all wiring and components under the chassis which were associated with the last IF amplifier power supply decoupling components i.e., R320, R321, C320 and C321, were placed on a small terminal board which was attached to the left side of the main chassis. Likewise the second detector and noise limiter components were placed on another small terminal board which was mounted underneath the chassis on the rear of the front panel. All connecting leads between these circuits and the rest of the receiver were replaced with shielded wire. In addition, the large molded capacitors originally used in these circuits were replaced with small mica or disk ceramic capacitors as appropriate. All of these capacitors were placed flat against the chassis and all leads were made as short as was possible.

2.6.4 Results of Changes: When these changes had been made, the suppression of pulse interference by the noise limiter was very pronounced, the improvement being of the order of 20 to 30 db over that obtained before the changes were made. Although these changes were made only to one model of the R361 receiver, it is felt that similar difficulties are taking place in all of these receivers which are of this same construction. A considerable improvement in the ability of these receivers to reject pulse interference can be obtained by making the simple changes listed in this note.

## 3. Conclusions and Recommendations

Several techniques have been presented which are effective in suppressing interference to communication and measurements equipment. High level pulse interference can be reduced or eliminated by the use of blanking techniques such as that employed in the pulse blanking equipment described in this report. At high signal levels an RF limiter is effective in improving the signal to interference ratio. However, successful operation of these limiters at signal levels below zero dbm was not obtained.

Several techniques are available for the reduction of CW interference to communications and measurements equipment. These techniques are: (1) highly selective passive preselectors, (2) low noise selective RF amplifiers, (3) fixed operating point balanced mixers, and (4) balanced local oscillators with low harmonic output. These devices when operated in combination have demonstrated that the spurious response rejection of currently used equipments can be greatly increased. As demonstrated by the adaptor described, this increase can be obtained by the conscientious application of well known principles.

It is recommended that the more promising techniques treated in this report be further developed to the point where they may be readily applied to field equipments with a minimum of modification.

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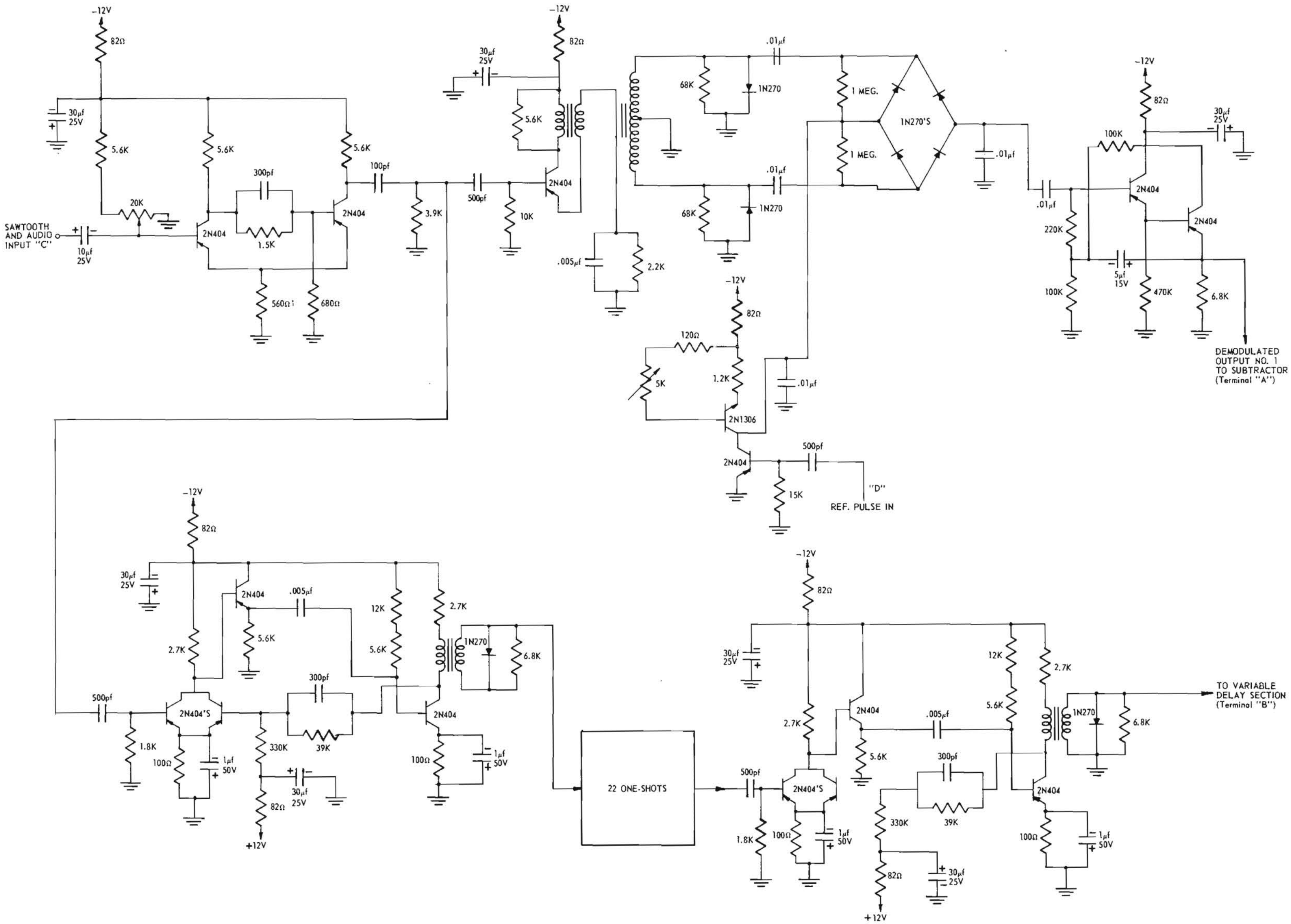
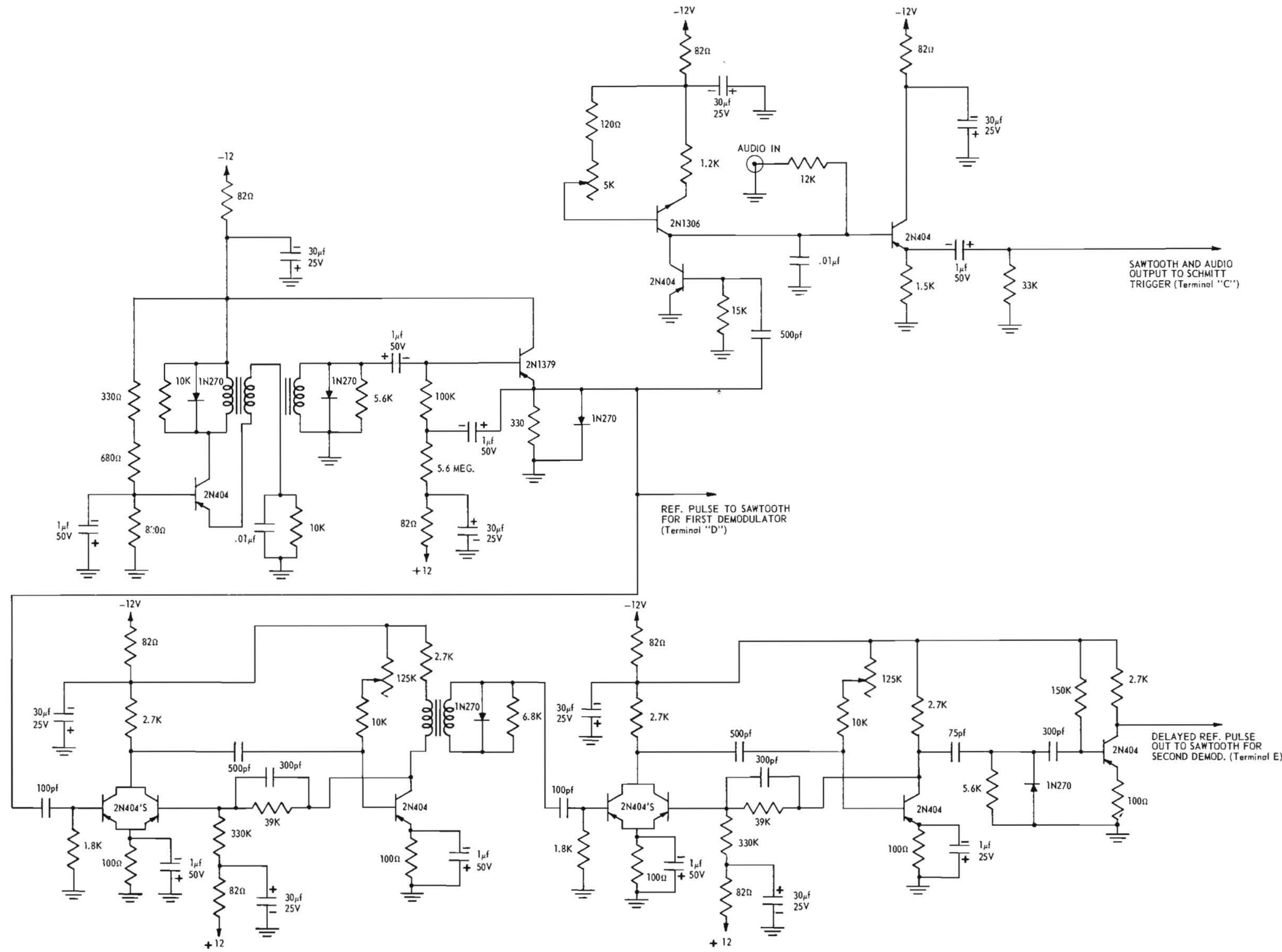
2. Robinette, S. L., "Pre-receiver Attenuator Reduces Intermodulation," Electronics, November 6, 1959

5. Appendix

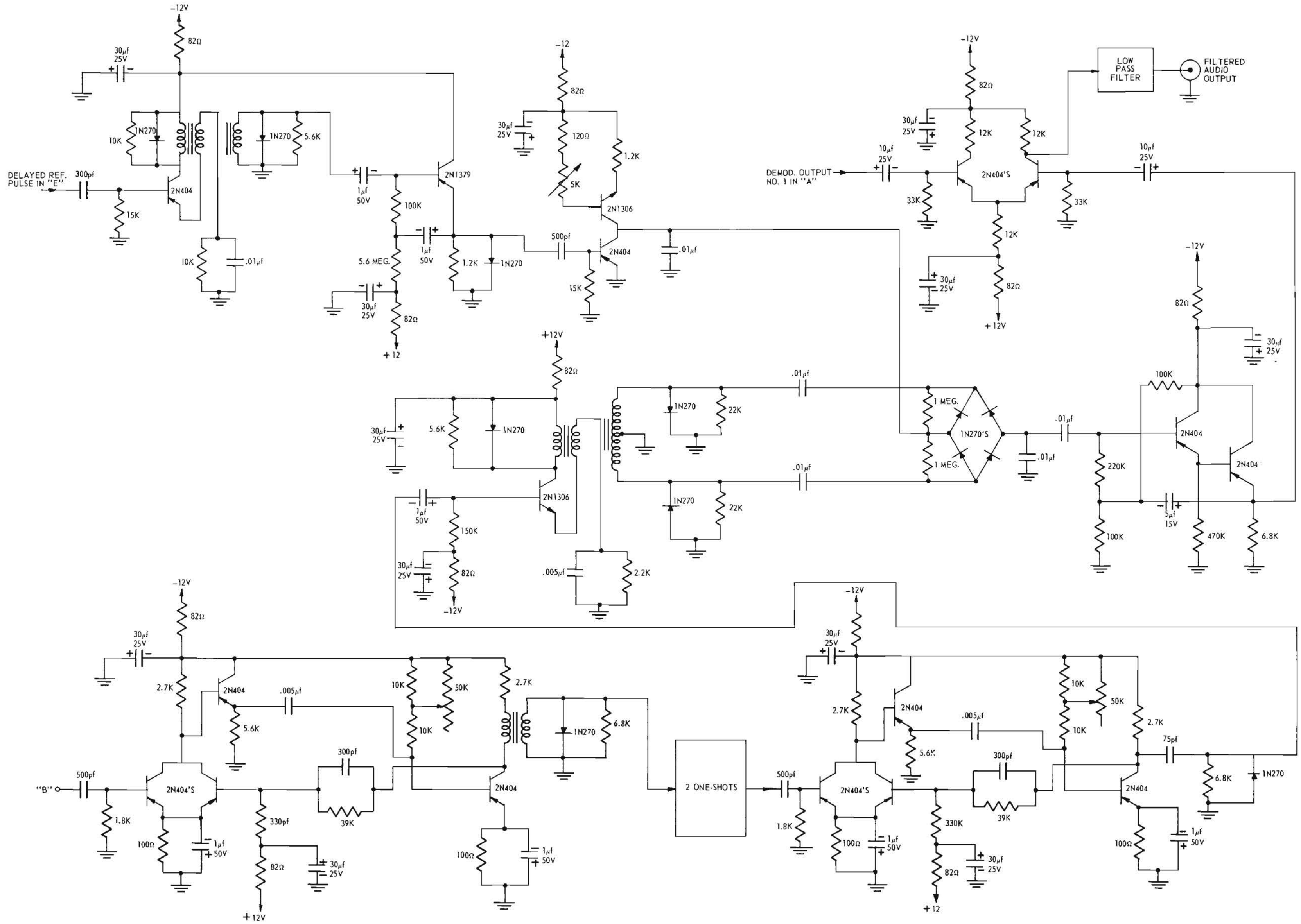
APPENDIX I

SCHEMATIC OF DELAY LINE FILTER


SAWTOOTH AND AUDIO INPUT "C"
2N404
1N270
1N270'S
2N1306
"D"
REF. PULSE IN
DEMODULATED OUTPUT NO. 1 TO SUBTRACTOR (Terminal "A")
2N404'S
22 ONE-SHOTS
TO VARIABLE DELAY SECTION (Terminal "B")
-12V
+12V



AUDIO IN
SAWTOOTH AND AUDIO OUTPUT TO SCHMITT TRIGGER (Terminal "C")
REF. PULSE TO SAWTOOTH FOR FIRST DEMODULATOR (Terminal "D")
DELAYED REF. PULSE OUT TO SAWTOOTH FOR SECOND DEMOD. (Terminal E)
2N1306
2N1379
2N404
2N404'S
1N270
-12V
+12


DELAYED REF. PULSE IN "E"
DEMOD. OUTPUT NO. 1 IN "A"
LOW PASS FILTER
FILTERED AUDIO OUTPUT
2 ONE-SHOTS
"B"
2N404
2N1379
2N1306
2N404'S
1N270
1N270'S
-12V
+12V